TECHNICAL REPORT RT-83-4

EXPERIMENTAL MEASUREMENT OF THE NEAR-FIELDS OF A UHF HORN ANTENNA

George R. Edlin Test and Evaluation Directorate US Army Missile Laboratory

October 1982



U.S. ARMY MISSILE COMMAND

Redstone Arsenal, Alabama

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The purpose of this report is to determine quantitatively some of the practical limitations of testing in the near field of UHF horns; in particular,			
how near to the horn aperture is the field reasonably uniform in amplitude and			
phase for a given size test object.	buy district in department of		
This report describes in detail near field me electric (E) and magnetic (H) vectors from the ape			

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a distance of 4.8 meters. These measurements encompass a rectangular volume of dimensions 2.4m x 3m x 4.8 m. The data spatial separation (1.a., spatial resolution) is 0.1m in each orthogonal direction.

The E and H field data is analyzed and displayed graphically. It is analyzed independently and combined to show the wave impedance behavior in the near zone.

By observation of the data presented, one can immediately develop engineering guidelines for defining the useful test volume in the near zone of a UNF horn. (Culture)

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I. INTRODUCTION

A. Background

The development of numerous high-power radio frequency (RF) emitters in the ultra-high-frequency (UHF) range has proceeded simultaneously with the continuing emergence of increasingly RF susceptible solid-state electronic components (particularly the most recently developed complementary metal-oxide-semiconductor (CMOS) technology). These concurrent technological developments have contributed greatly to increase the number of radio frequency interference (RFI) and electromagnetic compatibility (EMC) problems facing the communications engineers. It is assumed that these problems will loom even larger in the mid 1980's.

These RFI and EMC problems manifest themselves in two major categories when addressed in the particular domain of design, development, and testing of guided missile systems. The two categories are often referred to in the defense community as hazards (i.e., safety) and operational (i.e., functional dependability). In the first, the obvious area of major consideration is personnel safety. The protection and prevention of damage to system hardware and mission success potential are important, but secondary to personnel considerations. The safety problem area is dominated by careful attention to the electroexplosive devices (EED's) which are used to trigger missile launches, automatic eject mechanisms, and weapon detonations. The EED's are susceptible to RF energy and can be activated inadvertently by stray electromagnetic energy. Obviously, premature activation of these triggering systems may be extremely dangerous and is always undesirable. In the second category, system performance is the major consideration. Many external and internal factors influence system operational functions. One of these is inadvercent exposure to significant levels of electromagnetic energy at system critical frequencies (i.e., RFI). This RFI may cause system upset or malfunction. The RF energy levels required to produce undesirable effects in functional dependability are usually significantly lower (often several orders-ofmagnitude) than those encountered in hazards (safety) considerations.

Although the problems addressed are most relevant to guided missile systems, the ramifications are pertinent to any military and/or civilian communication and control system which exposes susceptible electronic circuits to potentially dangerous levels of electromagnetic energy in the UHF spectral regime.

Most of the undesirable effects mentioned above are directly related to the intensity of the spurious electromagnetic energy. As nature often seems to dictate, the intensity varies inversely with the distance (or some exponential power thereof) from the source. Consequently, the most serious environments for RFI and EMC effects are often in the near zone of radiating antenna structures. A survey of some pertinent documentation 1, 5, 6, 7 reveals continuing interest in the following areas:

- 1. Electromagnetic interference in electronic systems in the near field of high-gain microwave antennas.
- 2. Near field measurements to meet safety requirements.

- 3. Determination of far field patterns from near field measurements.
- 4. Mutual gain of multiple antennas in the near field.
- 5. Radar-cross-sections (RCS's) in the near zone.

An expanded discussion on some of these areas and referenced documents has already been published by this author and is included as Appendix A.

The area designated as number one is of particular interest to the US Army in testing the complex missile systems under development at the US Army Missile Command (MICOM).

As a necessary requirement in testing the complex US Army missile systems for potential RFI and EMC problems, each missile system must be exposed to electromagnetic energy levels with prescribed amplitude, frequency, polarization, and spatial field distribution characteristics. These are identical to or suitably simulate potential environments which the missile system might encounter in actual transportation and/or deployment. This requirement demands RF test capability which includes high-power sources (transmitters of several kilowatts) over a wide range of frequencies (100kHz-10GHz) coupled appropriately with broadband antenna systems (logperiodic, ridged horns, etc.) with selectable polarizations. As previously mentioned, the response of a system to the electromagnetic field imposed on it from an actual source antenna is of keen interest in both hazards and operational test requirements. In order to subject the test item or system to the proper field environment, two fundamental approaches are used. The first is quite straightforward: Fquipment to be tested is placed at an appropriate orientation and distance from the actual source antenna, and the spurious effects induced in the test system by the source are measured, recorded, and probable coupling mechanisms are identific. This approach is certainly valid; however, it is often very expensive and sometimes completely impossible from a logistical point of view. The second approach to the testing problem involves the concepts of simulation and/or extrapolation. The system to be tested is placed in one or more potentially "compromising" crientations with respect to a generic type broadband antenna and systematically illuminated with a broad range of frequencies, modulations, and polarizations. System response is again measured, recorded, and analyzed with particular emphasis on the nature of the response as a function of the electromagnetic field amplitude, frequency, polarization, and system orientation. From these general variations, extrapolations are carried out which hopefully predict how the system under test should react to the anticipated actual source of spurious electromagnetic radiation. Although this approach entertains many assumptions and a priori decisions about the test system behavior, it is effective in simulating a wide variety of environments. This approach allows some systems to be tested for which it would be virtually impossible to accomplish by using the first approach discussed. One of the major problem areas associated with the simulation/ extrapolation approach is a requirement of a high-intensity field level at all frequencies and polarizations. One of the newer techniques for establishing the required high-intensity field is the major subject of this report.

B. Statement of the Problem

This research report describes an effort to characterize quantitatively the near zone region of a high-gain UHF pyramidal horn antenna and subsequently to use this characterization to outline specific ways in which this information can be used to significantly enhance current technology in EMC/RFI testing.

C. Existing Solution to the Problem

There are two existing solutions to the problem of purforming high-level far rield testing. The first is to test at lower levels and extrapolate to the higher levels via analysis in conjunction with direct circuit injection. The analysis alone requires complex computations and still cannot account for nonlinear effects such as dielectric breakdown. The use of direct injection can solve the breakdown problem; however, the process is difficult to accomplish due to compact physical constraints and complex impedance matching problems, i.e., the coupling of the test signal can adversely affect the circuit under test.

The second way to accomplish the test is to have the necessary RF transmitting power available to produce the high field levels required at the recommended far field test distance. The normally accepted far-field test distance from the source antenna is given by $2D^2/\lambda$ where D is the largest dimension of the source anterna aperture and λ is the wavelength of the transmitted signal. This generalized formula has been cited in the Navy report, "Near Field and Fresnel Zone Directive Antenna Coupling Program Final Quarterly Report," report #6174D, February 1973. This criterion represents the distance at which there is a reasonably small phase shift across the test volume. In addition, the theoretical wave impedance is approximately 377 ohms. Satisfying these conditions causes the near some region to extend many meters from the transmitting antenna. The problems with the second approach are the technical and financial constraints. Broadband amplifiers sell for around \$50K each, and the band coverage is approximately 200 MHz each from 200 MHz to 1000 MHz. Even these transmitters cannot simulate multikilowatt radar units. Consequently, the test engineer must often forget about continuous frequency coverage and depend on multikilowatt narrow band transmitters. The cost of these units may run several \$100K. This approach requires millions of dollars to build a high-level facility. Although technically feasible, this is a very expensive way to accomplish the required objectives.

D. Proposed Solution

The approach this work explores is the utilization of the near some of large, high-gain transmitting antennas. The use of the near some requires that the test volume be completely characterized electromagnetically. The methodology of this investigation utilizes accurate spatial resolution amplitude measurements of the three orthogonal components of both the E and H field intensities. The development of small nonperturbing E and H field probes concurrent with high-speed automated data systems has made it possible to make and record the large number of data points necessary to electromagnetically characterize the desired test volume. These experimental

measurements are compared to α standard waveguide type aperture model. These results are utilized to develop the wave impedance value as a function of distance from the aperture. The E and H field characteristics are defined in the near sone, and a corrected antenna gain expression is formulated to allow calculation of field strengths for distances as close as $D^2/4\lambda$ from the radiating aperture. At the $D^2/4\lambda$ distance in front of the UHF horn under consideration, a 1-kW transmitter produces the same electromagnetic field strength as a 20-kW transmitter at the recommended $2D^2/\lambda$ distance from the UHF horn aperture.

II. TECHNICAL APPROACH

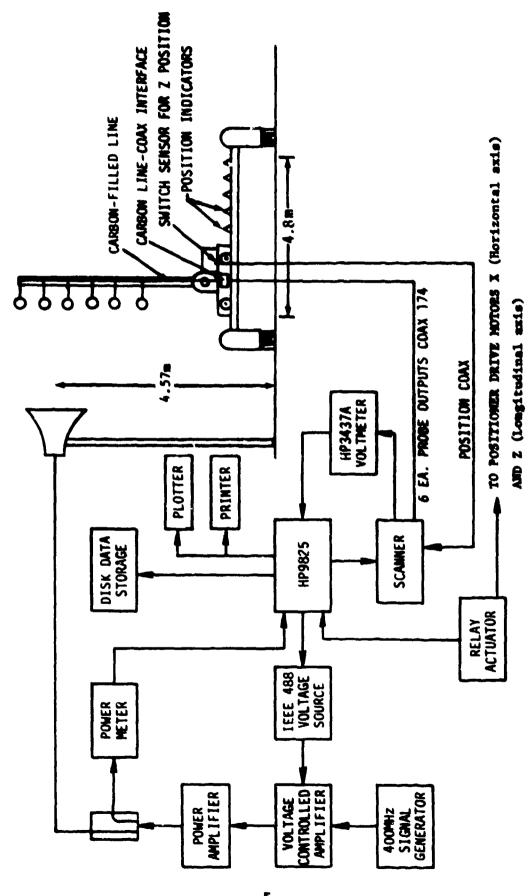
A. Summery of Technical Approach

The technical approach used for this report is to develop suitable non-interfering, polarization selective, E and H sensors to accurately measure the electromagnetic fields. The next step is to construct a means to spatially position the sensors. An additional requirement is to develop a data acquisition system to read, record, and spatially relate each data point. The final step in this sequence is to process the data and compare the measured data to the theoretical models.

B. Development of the Measurement Techniques

Probe Description - The probes utilized for the characterization of the three orthogonal components of both the E and H fields are polarization selective, pseudo transparent, non-interfering type sensors. Both the E and H sensors use diode detection with carbon filled lines to connect the sensor to a 100 megohm input impedance voltmeter. The H field sensor is a double gapped loop wired electrically to cancel any component of the E field which drives the output gaps. The E field probe is an electrically short dipole which is also diode-detected and interfaced to the same high impedance voltmeter via carbon filled lines. The calibration of the E and H sensors is accomplished by establishing known E and H fields in a Crawford cell. The Crawford cell is an expanded coaxial line which can be used to establish accurate fields. The sensors are placed in known fields and the detected outputs measured and related to the standard field levels. The calibration curves developed are used to relate the raw data to actual field strengths. Appendix B contains complete explanations of the design and calibration of the E and H field probes.

Three-Dimensional Scant...: - The three-dimensional scanner is constructed using 6 inch polyvyn'chloride (PVC) pipe for the base section as shown in Figure 1. The trolley which moves the probes in the Z direction (toward and away from the antenna aperture) and the vertical support pole are also PVC. The position is indicated to the computer via a microswitch and a 9 volt battery in series with the carbon filled lines. The trip positions are located at precise 10 cm intervals along the Z axis from 0.1 to 4.8 meters away from the aperture. The X position is also transmitted to the HP 9825 controller by the X position (horizontal position) microswitch. The trips for the X position are also 10 cm apart for -1.5m<X<1.5m. Six E or six H probes are mounted 40 cm apart on the vertical support with provision for lowering the support in 10 cm increments three times. This gives



Pigure 1. Data acquisition system.

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complete coverage for 1.2m above the antenna center line to 1.2m $(-1.2m\le Y\le 1.2m)$ below the antenna center line. More detailed information on the positioner is included in Appendix C and in the discussion of the data acquisition system.

Data Acquisition System - The data acquisition system consists of a standard gain horn (gain 18 dB nominal, frequency, 300 to 550 MHz) mounted in the vertical F field polarization orientation with the center line 4.57m above the ground. The computer controller is an HP 9825. The excitation for the horn is a Watkins-Johnson (WJ) synthesizer in series with a voltage controlled attenuator. This allows the computer to adjust the excitation level to the solid-state power amplifier. The power is monitored using a directional coupler and an IEEE 488 bus operated HP 436 power meter. In this way the controller is able to monitor and set the power level before each scan. The frequency used is a constant 400 MHz. This frequency is compatible with the horn and field measurement system. The 4.8m probe scanner, along wit the HP 3495A scanner, HP 3447A system voltmeter, HP 9825 controller, and 8-inch floppy disk comprise the major hardware components utilized in the data collection system. Figure 1 presents the block diagram for this system. The probe positioner configuration allows for acquiring data precisely every 10 cm for each of the three orthogonal directions. The data measurement matrix results in 37,200 locations for each of six orthogonal components.

The total data system is controlled by the HF 9825 controller. The operator has only to turn on the equipment, set the frequency of the exciter to 400 MHz, and position the mechanical probe positioner to the starting position. The operator then executes the control program and inputs the type probe (E or H) and the component of the field being measured (vertical, horizontal or radial). The operator inserts the proper disk and starts the data acquisition process. The control system automatically adjusts the power to 20 watts into the transmitting system and starts the first scan which is in the Z axis (along the antenna radiation axis). The trip points located at 10 cm intervals along the Z direction synchronize the data readings of each of the six probes mounted on the vertical support. The program reads the raw data and stores it along with the X (horizontal), Y (vertical) and Z (radial) position information after applying proper calibration factors. After the data storage is completed, the positioner moves over 10 cm in the X direction and makes another scan in the Z direction. This process is continued until all the X positions are measured. The vertical probe support is then vertically shifted by 10 cm (Y axis), and the original process is repeated for a total of four times to complete one orthogonal component of the E or H field. A detailed description of the positioner is contained in Appendix C. Each of the orthogonal components are stored on separate 8-inch floppy disks. One complete set of data is 223,200 data points contained on six disks. These data are also printed as they are taken to allow for realtime observation as well as to check for possible disk storage problems.

C. Data Collection

w Data - The E field and H field probes have DC outputs on the order of aillivolts. The diode detector is a nonlinear device; therefore, the polynomial that relates the field amplitude to the voltage output of the

probe is complex and is of the form $F_L = C_1V_0 + C_2V_0 + C_3V_0$. Where F_L is the field level, V_0 is the probe output, and C_1 , C_2 , and C_3 are the calibration coefficients determined by putting the probe in known electromagnetic fields at several levels and using a third order polynomial regression program. This polynomial will allow the computer to rapidly compute the correct field level for any probe output voltage.

Data Processing - To convert the data to meaningful information, the potential reading must be multiplied by the appropriate polynomial and stored in data arrays on the disk. Each disk contains a single orthogonal component of the E field or H field. The orthogonal components of the E and H field can then be printed in tables or plotted in two-dimensional or three-dimensional graphs to be analyzed or compared to theoretical plots.

D. Comparison of Collected Data to Theoretical Models

General Discussion of Measured Data - Observation of the data yields some very significant information to the microwave test engineer. First, if one considers the impedance curve shown in Figure 2, one notices that the wave impedance does not vary significantly from the 377 ohm far field values, even as near to the aperture as one meter. This implies that only the E field has to be measured to characterize the power density at test points very close to the antenna aperture. The measured data showing the wave impedance is very well behaved as close as $D^2/4\lambda$ from the aperture. The centerline data plotted in Figure 3 shows that the vertical E field is also well behaved as near as $D^2/2\lambda$ and is reasonably good at distances greater than $D^2/4\lambda$ from the aperture.

Aperture Distribution Model - The analytical model for this was TE₁₀ mode amplitude (cosine in X) distribution with a quadratic phase term.

The aperture distribution was used with the vector Smythe-Kirchhoff³ approximation for diffraction to solve for the spatial representation of the radiated field. This relationship is given in equation (1).

$$\vec{E}(x,y,z) = \frac{1}{2\pi} \nabla x \int (\hat{n} X \hat{E}) \frac{e^{jkR}}{R} da'. \qquad (1)$$

Figure 4 shows the horn geometry and aperture associated coordinate system. The TE_{10} waveguide E-Field distribution is given by equation (2).

$$\vec{E}(x,y,0) = E_0 \cos \frac{\pi x}{a} \exp \left[-jk\left(\frac{x^2}{2l_H} + \frac{y^2}{2l_E} + \alpha\right)\right] \hat{U}_y . \quad (2)$$

Substituting the expression in equation (2) into equation (1) and solving for the associated E field and H field components we have equations (3) thru (9).

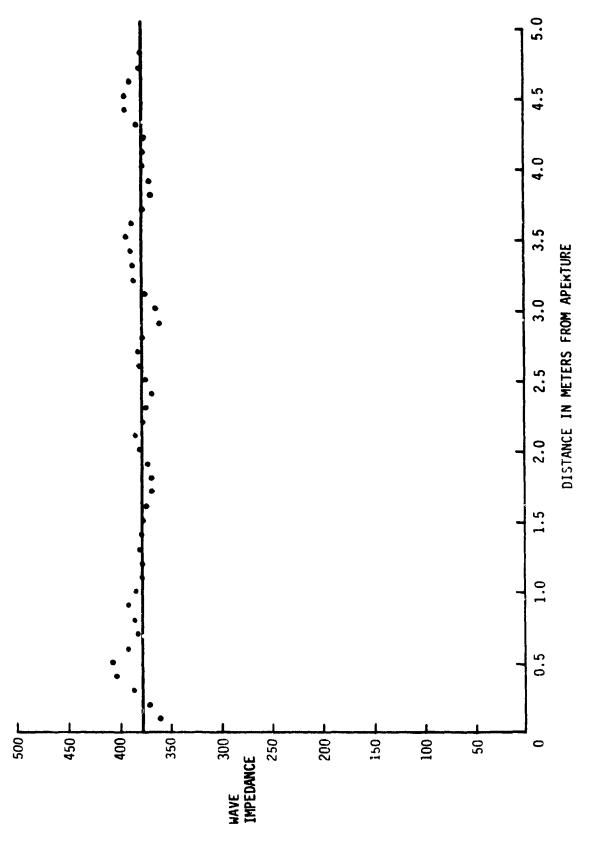


Figure 2. Wave impedance $^{2}_{o(0)tms}$, distance (2) (meters) from the aperture.

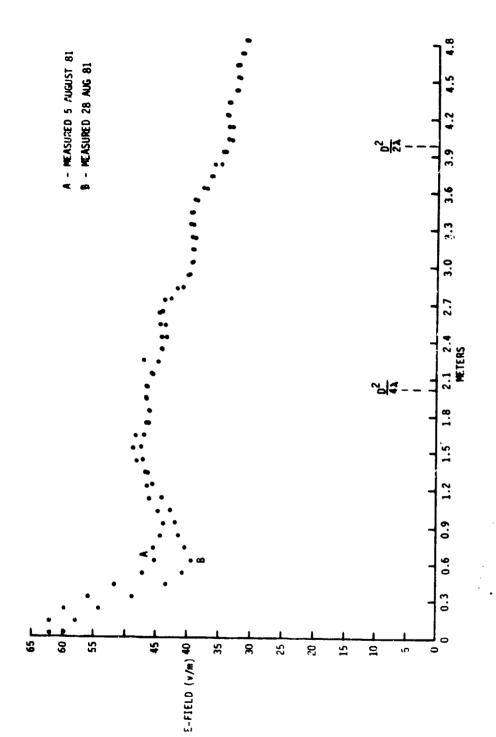


Figure 3. Vertical E-field vs. 2 from the Aperture.

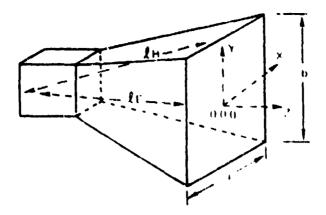


Figure 4. Geometry and coordinate system of UHF pyramidal horn.

$$E_{y} = \frac{E_{0}}{2\pi} \int \int \cos \frac{\pi x'}{a} \frac{e^{-jkB} z (1-jkR)}{R^{3}} dx' dy'$$
 (3)

$$E_{Z} = \frac{E_{0}}{2^{\pi}} \iint_{Aper} \cos \frac{\pi x'}{a} \frac{e^{-jkB}(y' - y) (1 - jkR)}{R^{3}} dx'dy'$$
 (4)

where

$$R = \left[z^2 + (x - x^*)^2 + (y - y^*)^2\right] \frac{1}{2}$$
 (5)

$$B = \left[\frac{x^{*2}}{2^{\ell}_{H}} + \frac{y^{*2}}{2^{\ell}_{E}} - R \right]$$
 (6)

$$H_{x} = \frac{jE_{0}}{2\pi\omega\mu} \int \int \cos\frac{\pi_{x'}}{a} \frac{e^{jkB}}{R^{3}} \left\{ \left(k^{2} - \frac{3}{R^{2}} + 3j\frac{k}{R}\right) \left[(y-y')^{2} - Z^{2} \right] \right\}$$

$$+\left(1-jkR\right)\right\}dx'dy'$$

$$H_{y} = \frac{jE_{0}}{2\pi\omega\mu} \int_{Aper} \int_{Aper} \cos\frac{\pi x'}{a} \frac{e^{-jkB}}{R^{3}} (y-y') (x-x') \left(k^{2} - \frac{3}{R^{2}} + \frac{13k}{R}\right) dx'dy'$$
 (8)

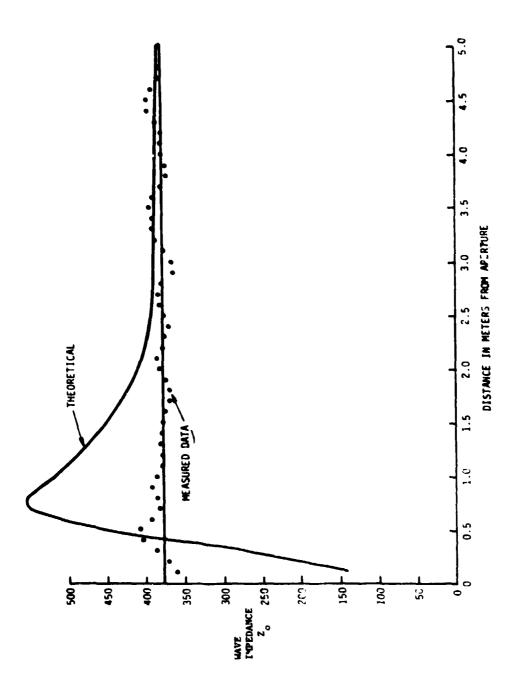
$$H_{z} = \frac{jE_{0}}{2\pi\omega\mu} \int \int \int \cos\frac{\pi x'}{a} \frac{e^{-jkB}}{R^{3}} z (x-x') \left(k^{2} - \frac{3}{R^{2}} + \frac{j3k}{R}\right) dx'dy'.$$
 (9)

This model has been utilized by several investigators^{2, 4}. A detailed derivation of the commonly accepted analytical model is presented in Appendix D, and numerical evaluations and presentation of the model are included in Appendix E.

Figure 5 compares the measured wave impedance as a function of Z with that obtained theoretically in Appendix E. The measured data do not exhibit nearly as sharp a rise as the computed curve, although it does show a similar shape with perturbations around the 377 ohm line as the distance in the Z direction exceeds 0.75 meters. These small perturbations from 0.75 to 5 meters may be the result of standing waves which seem to be visible in the three-dimensional plot (Figure 6). The fact that the impedance of the measured data does not peak as high or dip as low as the model can be attributed to four main factors: (1) inaccuracies in the model, (2) inaccuracies in the horn construction, (3) spatial integration of the E field amplitudes that change rapidly with Z produced by finite probe size, and (4) standing waves causing constructive and destructive interference in the field amplitudes.

The significance of Figure 5 is two-fold. The first and main point is that the near field wave impedance is essentially the same as far field wave impedance to distances from the aperture of the test horn antenna as near as one meter. In terms of the wavelength of the experimental setup, $D^2/8\lambda$ equals approximately one meter. Secondly, the results of Figure 5 imply that only one of the field components needs to be measured to establish the power density of the radiation field. This fact makes it feasible to utilize the near field for testing purposes. A plot of the analytical model with the measured data as a function of the discance Z (meters) from the aperture is presented in Figure 7. It is very obvious that the measured data do not increase as fast as the model and do not oscillate as much as the model inside the 1 meter or $D^2/8\lambda$ distance from the aperture. This makes the field in the very near zone $D^2/8\lambda$ far more usable than was predicted by the model.

Multipole Model - If one looks closely at Figure 8 which has the 1/r relationship plotted along with the measured data, it is obvious that the measured data do not increase nearly as fast as the 1/r curve near the aperture. This fact also contributes to the usefulness of the volume near the aperture for testing of small systems, i.e., the rate of change of the E field is not appreciably different at points 2-3 meters from the aperture



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Figure 5. Wave impedance $\left(\frac{Ey}{Hx}\right)$ vs. Z for rectangular pyramidal horn.

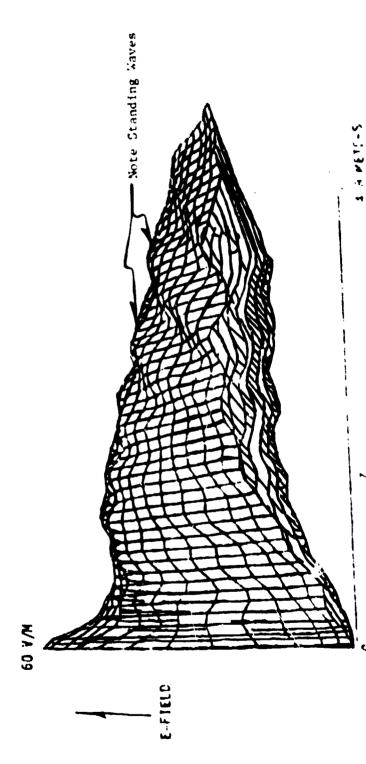


Figure 6. 3-Dimensional plot of E-field (V/m) vs. horizontal position and radial position (side view).

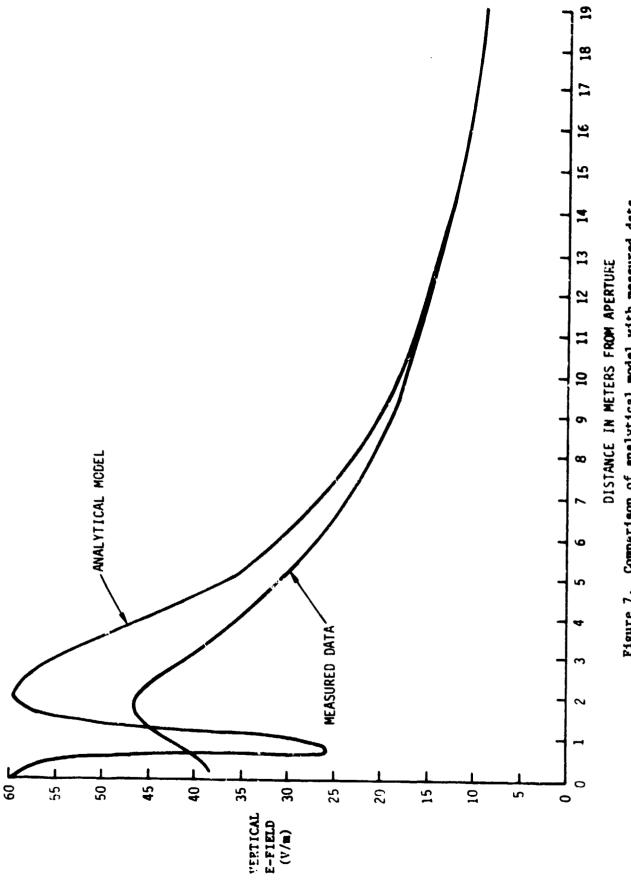
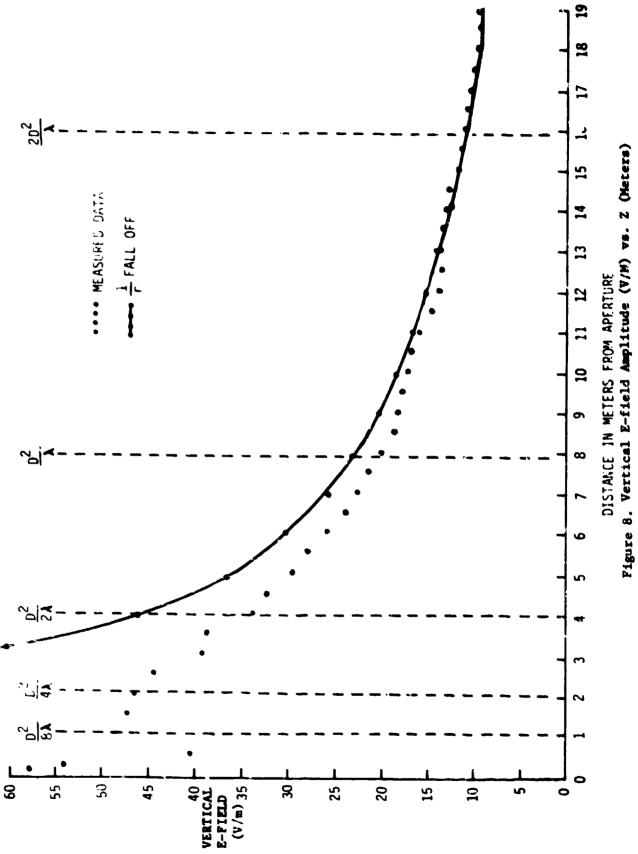


Figure 7. Comparison of analytical model with measured data.

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and points 6-8 meters from the aperture. The advantage of testing in the near zone is very obvious; if one moves in to half the distance, in general, the field strength is effectively doubled, which means to have produced the same field strength at the greater distance would have required a power increase of four times. This simple illustration then implies that for a small system, the test engineer can utilize the near zone, $D^2/2\lambda$, and effectively achieve the same test field with a 1 kW transmitter that would require a 16 kW transmitter at the classical $2D^2/\lambda$ position. It is apparent that the 1/r model is not accurate at radial distances closer than D^2/λ . A useful model which is accurate at a closer distance is the multipole model. The form of the electric field for the multipole model is given as

$$E = \left(\frac{A_1}{r} + \frac{A_2}{r^2} + \frac{A_3}{r^3}\right) (P_T)^{1/2}$$
 (1)

where

E = E field in volts per meter (V/m)P_T = power of the transmitter (Watts)

r = distance from the aperture in meters

A₁, A₂, A₃ = experimental constants

A comparison of the multipole model to the experimental data is shown in Figure 9. This model can be used as near as $D^2/4\lambda$ with an accuracy of approximately +0.5 dB. Details of the derivation of this model are presented in Appendix F.

Engineering Model - Since the main objective of this work is to develop a methodology for the utilization of the near zone volume for EM testing, one of the main considerations is to develop an engineering model which would predict the E field at any frequency and radial distance from the antenna.

The E field in the near zone of a UNF antenna predicted by the engineering model is given by equation (2)

$$E = \left\{ \left[\frac{G(1 - \exp(-4r/(2D^2/\lambda)))}{4\pi r^2} \right] 377P_T \right\}^{1/2}$$
 (2)

where

E = E-field (in V/m)

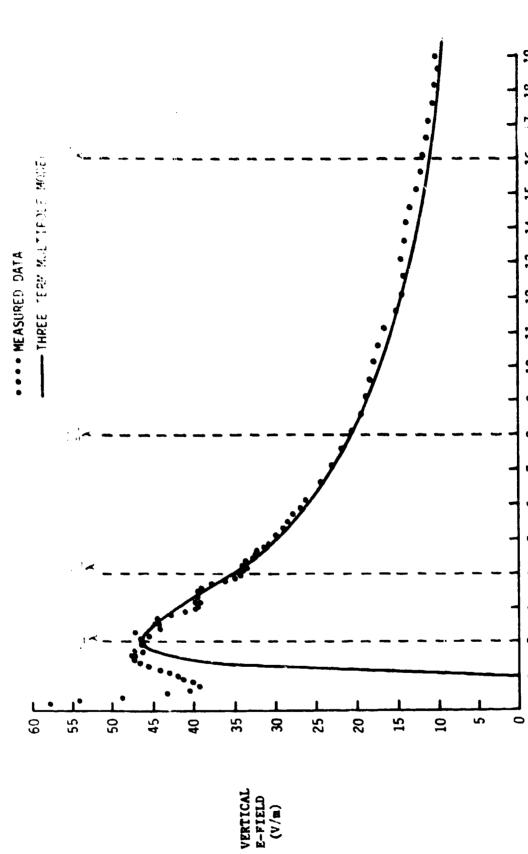
r = distance (in meters) from the aperture

D = largest dimension of aperture (in meters)

 λ = wavelength (in meters)

P = transmitter power (in Watts)

G = far field gain (absolute, not decibels)



Pigure 9. Multipole predicted E and measured data vs. Z.

DISTANCE IN METERS FROM APERTURE

For additional detail of the engineering model refer to Appendix G. The comparison of the engineering model with the measured data is shown in Figure 10. This model will predict the E field of the test horn in as near as $D^2/4\lambda$ with accuracy of approximately +1.5 dB.

Three-Dimensional Data Plots - One of the best ways to show the radial uniformity of vertical E field (E) and horizontal H field (H) is to observe a few 3-D plots. The 3-D plots are presented with the aperture of the horn on the observer's left; with X, the horizontal dimension; field amplitude, the vertical dimension; and Z, the radial dimension from the aperture increasing to the observer's right.

The vertical dimension displays either the amplitude of the E or H field at a selected Y position. This position is always held constant at sero which is center line for the horn for Figures 11 and 12 (see Appendix H). The observation of these figures gives one a visual presentation of the E field shape as a function of X and Z on the center line of the horn from the horn aperture to a separation distance (Z) of 4.8 meters. The intersection points of the cross hatched lines represent the measured data points. These figures are given to show shape and are not scaled. Appendix H contains all plotted data. Explanation of the data correlation is also contained in H. Careful observation of Figures 11 and 12 shows the possible effect of standing waves on E field amplitude. The standing waves referred to are the small perturbations in amplitudes along the Z direction. These waves give rise to some shifting of the wave impedance Z from the far field nominal value of 377 ohms.

The view in Figure 13 shows how the cosine distribution at the aperture diminishes as the radial distance increases. This figure is presented from the same aspect as Figures 14, 15 and 16 which allows the direct comparison of Figures 13 and 14. This comparison gives visual confirmation that the ratio of E to H is consistent across the X dimension of the test volume. Figures 15 and 16 show that the radial H and vertical H are consistent with the radial E and horizontal E shown in Figures 11 and 12, i.e., the vertical H compares with the horizontal E near zero, and the radial amplitude is low and falls off rapidly with increasing distance from the aperture.

III. CONCLUSIONS

A. Near Field Characteristics of the UHF Horn Antenna

Radial Uniformity of the Wave Impedance - The previously discussed Figure 3 shows the radial uniformity of the wave impedance along the center line of the antenna. The maximum deviation is approximately 33 ohms. This occurs near 0.5 meters and is less than 9% different from the far zone value.

The importance of this uniformity cannot be overstressed due to the simplification of the field measurement problem and the understanding of its interaction with various electronic circuits.

Radial Attenuation of the Nonradiative Components - An informative way to view the rapid fall off of the nonradiative component is to observe Figure 12 which shows the 3-D comparison of the radial E field, $\mathbf{E}_{\mathbf{p}}$, with the

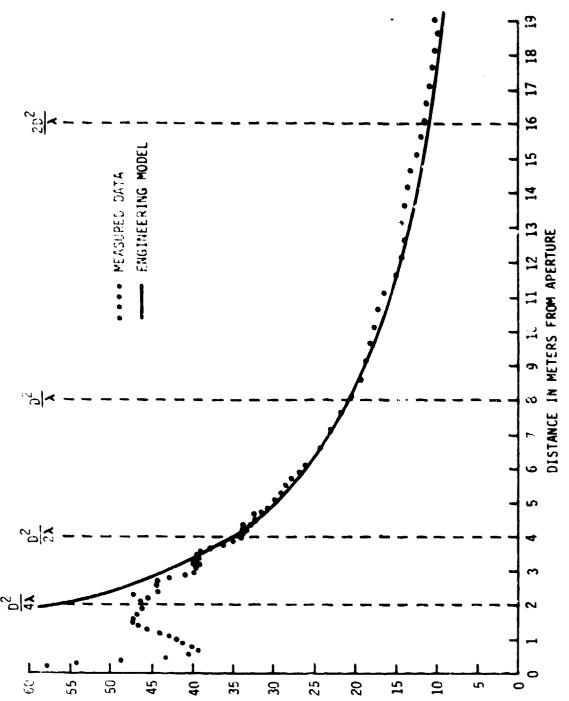


Figure 10. Engineering model compared with measured data.

VERTICAL E-VIELD (V'm)

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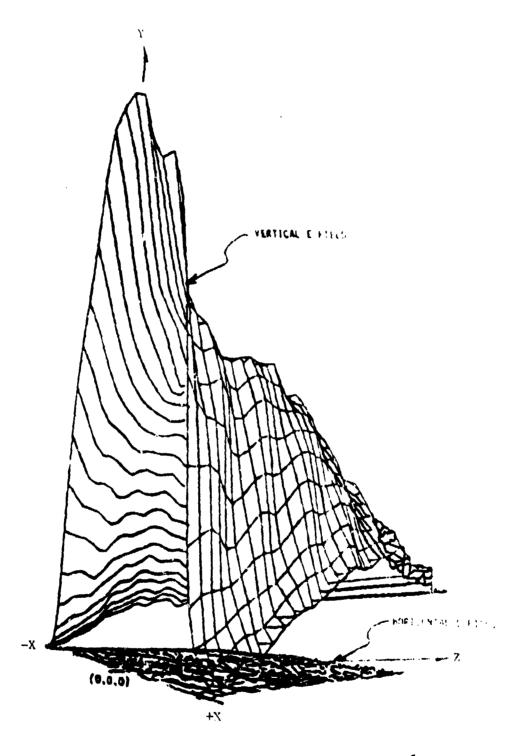


Figure 11. E_V and E_H amplitudes vs. horizontal position (X) and radial position (Z)

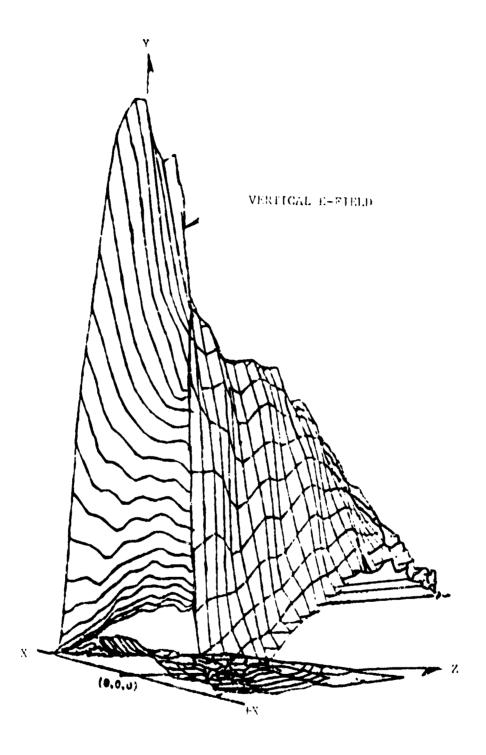


Figure 12. E_{V} and E_{R} amplitude vs. horizontal position (X), radial position (Z).

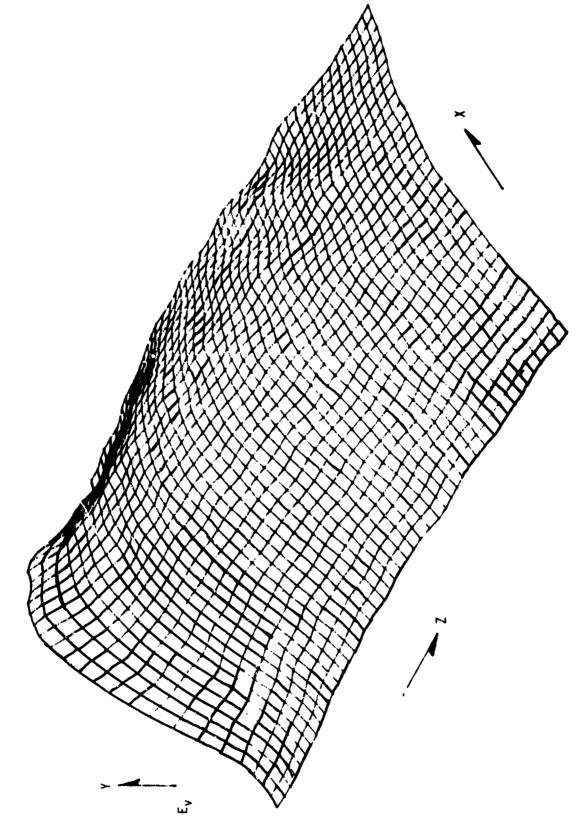
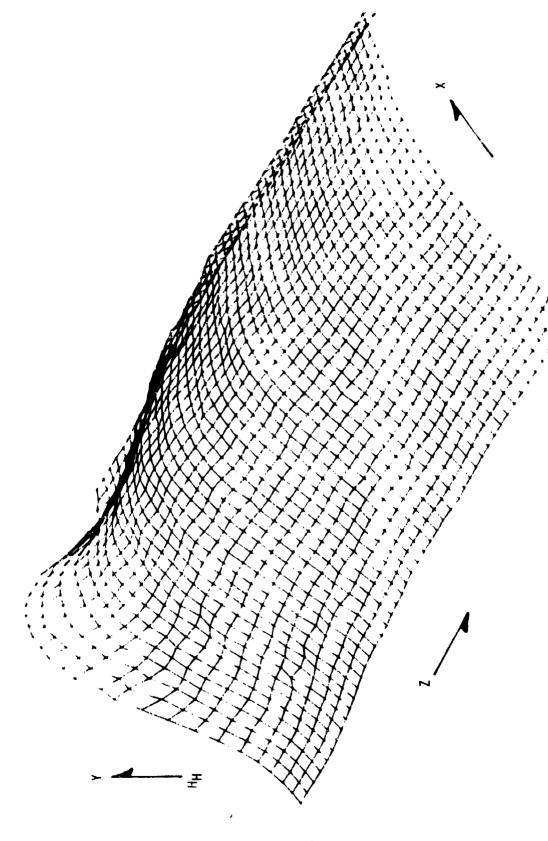


Figure 13.



 $H_{\rm H}$ amplitude vs. horizontal position (X) and radial position (Z). Figure 14.

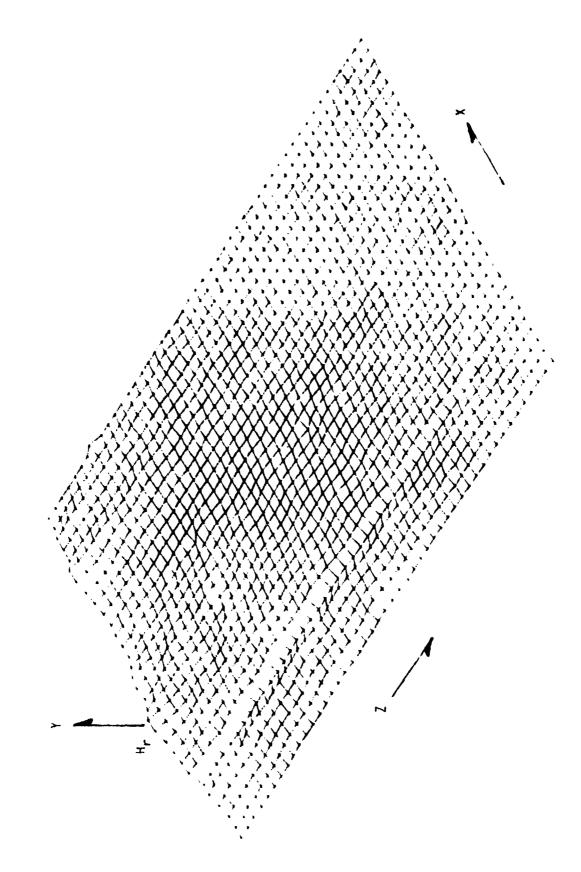
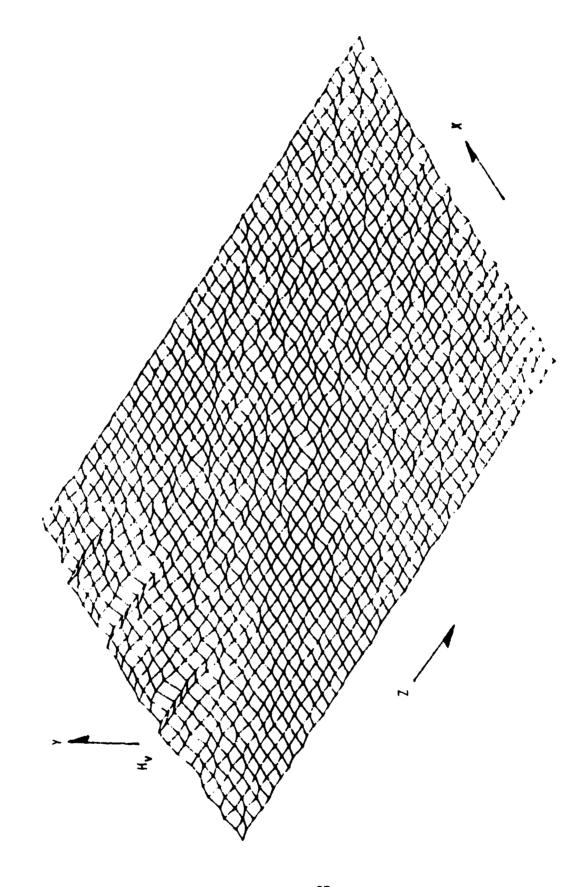


Figure 15. H_{R} amplitude vs. horizontal position (X) and radial position (Z).



H amplitude vs. horizontal position (X) and radial position (2). $_{\nu}$ Figure 15.

vertical E field, E_v . The E_R is 30 dB below E_v by the two-meter point. The horizontal E field, E_H , is compared to E_v in Figure 11. It is apparent that the E_H is approximately zero. Therefore, these nonradiative components cease to be significant unless the field observation or testing is to be carried out extremely close to the source aperture.

B. Impact on Practical Electromagnetic Testing Techniques

In conclusion, the objective of this work is to characterize the near field volume of a high gain UHF horn for use in electromagnetic testing. This task has been accomplished and the results used to develop a simple engineering model to predict the E field at any radial distance from $D^2/4\lambda$ to the far field interface, at $2D^2/\lambda$. Given a small test object such as a VIPER missile, one could run an electromagnetic hazards test as close as $D^2/4\lambda$ to the aperture. This allows a 1 kW amplifier to produce approximately the same field strength as a 20 kW amplifier at $2D^2/\lambda$, the classical distance for an acceptable field test. Consequently, for the horn used in this work, a small test item could be as close as two meters to the aperture.

The previous example represents the extreme practical limits for the utilization of the near zone test volume. As the test specimens increase in size, the test volume moves out to an appropriate position to produce a more uniform illumination of the test object. It is not necessary for the microwave test engineer to arbitrarily move to $2D^2/\lambda$ to run all of his EMI tests. This work demonstrates that both the E and H fields are well behaved in the near zone of standard gain horns. This allows high level sweep frequency testing which is not possible using the far field distance criteria of $2D^2/\lambda$.

C. Suggestions for Future Research

The process of obtaining and analyzing the data presented in this work has answered some of the basic questions facing the microwave test engineer who wishes to utilize the near zone of a UHF horn antenna. There are, however, other questions and problem areas which need to be explored in detail. One of the most difficult areas of consideration is that of the phase relationship of the E and H field components. To explore the phase relationships one would need to develop non-perturbing E and H field probes that would preserve the phase relationships of the E and H field components.

Another area that warrants consideration is the validation of near field test result, i.e., the direct comparison of specific electromagnetic radiation (EMR) results obtained in near zones with the results observed in far zone tests, thus confirming the validity of testing in the near zone of UHF horns. The validation should address the test object/antenna interactions, i.e., the effects of geometry, distance from aperture, and test object size relative to the test volume. In addition, near field testing techniques need to be extended to other antenna types such as the log periodic and ridged horns.

Development of probe calibration techniques to improve the accuracy of the E and H field measurements would greatly benefit EMR testing. It is also highly desirable to develop automated probe calibration techniques since

this would result in a considerable time savings as well as potentially improve the accuracy of probe calibrations.

Additional efforts are needed to validate new and innovative phase controlled multichamber testing as described by Riley, Patent Disclosure No. 4,255,750. The techniques suggested by Riley would greatly simplify testing systems with long interconnecting cables. This disclosure describes, in detail, the technique that needs to be developed and verified.

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APPENDIX A

A SURVEY OF ADVANCES IN MICROWAVE NEAR-FIELD MEASUREMENT

Reprint of MICOM Technical Report RT-80-9, June 1980, George R. Edlin.

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I. INTRODUCTION

The development of many high power emitters in the microwave frequency range has led to many problems which cannot be solved using classical far field techniques. The problems of particular interest may be grouped into the following areas:

- 1. Electromagnetic interference in electronic systems in the near field of high gain microwave antennas.
 - 2. Near field measurements for use in determining far field patterns.
 - 3. Mutual gain characteristics of multiple antennas in the near field.
 - 4. Near field measurements for occupational safety.
 - 5. Determination of Radar Cross Sections (RCS) in the near zone.

Interest in electromagnetic interferences in the Fresnel zone has increased in the past decade due to development of large phased array radars and high gain satellite antennas. Since these antennas are very large with respect to the wavelengths at which they are operated, the far field can extend a great distance from the antenna. The normally accepted distance is given by $2D^2/\lambda$, where D is the largest dimension of the antenna, and λ is the wavelength of the transmitted signal.

It is readily obvious that for antenna sizes in meters and wavelengths in centimeters, the near field zone extends many meters from the transmitting antenna. This causes a large number of electronic systems to be exposed to high level near field electromagnetic radiation. Many times this high level radiation causes circuit malfunction in some solid state systems, either by induction of extraneous signals into the system or by burnout of critical components. This situation has created a need for high level electromagnetic testing in both near and far fields. Since high levels are much easier to obtain in the Fresnel zone, it is desirable to relate the testing accomplished in the Fresnel zone to the appropriate far zone. Two major objectives could be met with this technique. First, only one near field test would need to be conducted. Secondly, by being near the transmitting antenna, the power requirements for producing high level fields would be reduced significantly. This would result in considerable cost savings on the required broad-band transmitters.

The need for determining far field patterns from near field measurements is again due to the development of large higher gain microwave antennas. The testing of these antennas requires very long, unobstructed antenna ranges to obtain accurate patterns and gains. Costs of acquiring and developing such ranges are prohibitive, therefore making development of near field pattern and gain techniques more desirable.

The use of multiple antennas located near each other has created an interest in mutual gain characteristics of microwave antennas. These situations are major problems for large naval ships, Army missile systems, and Air Force systems.

Occupational health considerations have also stimulated a considerable amount of research in the Fresnel zone. This work is necessary to determine how far the hazardous levels of electromagnetic radiation extend from the transmitting antennas.

II. REVIEW OF PREVIOUSLY PUBLISHED MATERIAL

Following is a review of material published in journals, reports, periodicals, and books in the past decade.

A. Mutual Gain Statistics for Microwave Antennas

The investigation of mutual gain of microwave antennas in the Fresnel zone is of great importance in any situation where multiple antennas are required. The work by researchers at Georgia Institute of Technology presents measurement results and analysis of the Fresnel zone statistical mutual gain characteristics of four C- and X-band radar antennas. The tests were conducted at ranges of 50 feet and 9 feet. The 9-foot distance was just short of touching due to antenna sizes.

Studies at Georgia Tech have shown that the minor lobe statistical gain characteristics for the horizontal plane patterns are described approximately by the Gaussian cumulative gain distribution curve for ranges well into the Fresnel zone. Some problems occur at very short Fresnel zone distances due to the broadening of the main lobe of the antenna, causing the cumulative gain distribution curves to deviate from the Gaussian type curves.

Previous mutual gain experiments have shown that if the statistical gain distributions of two antennas are approximately Gaussian, this statistical mutual gain distribution of the two antennas is approximately Gaussian. For these tests the shortest separation was 50 feet and no site-effect objects were included. Due to the questions concerning coupling at very short Fresnel distances, two tests were conducted at 50 feet and x-band tests were conducted at 9 feet.

From graphical representation reproduced from this report (Figures 1 and 2), it can be noted that at a range of 50 feet, the measured points and predicted points agree very well. This indicates that if the statistics of the individual antennas are known accurately, the mutual coupling can be predicted very accurately.

The curves at the 9-foot range do not match as well as those at the 50-foot range. The disparity occurs at the top of the curves. This is the result of the main beams broadening near the antennas and thus deviating from Gaussian distributions.

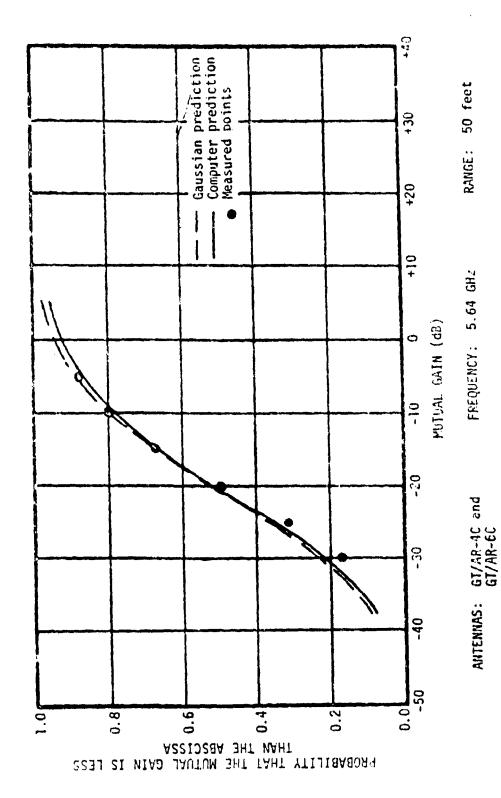
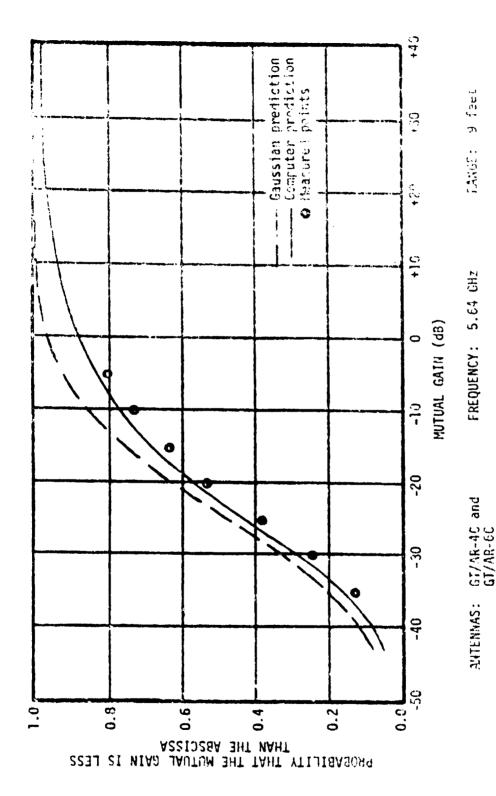


Figure 1. Measured and predicted cumulative distributions of the mutual gain of the GT/AR-4C and GT/AR-6C antennas for the clear site test at a (Extracted from Reference 1.) Fresnel zone range of 50 feet.

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Figure 2. Measured and predicted cumulative distributions of the mutual gain of the GT/AR-4C and GT/AR-6C antennas for the clear site test at a (Extracted from Reference 1.) Fresnel zone range of 9 feet.

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Although the Gaussian predictions at very short fresnel zone ranges do deviate from the measured data points, in many cases the predictions are sufficiently accurate to satisfy the requirements. If the Gaussian predictions are not accurate enough, then the numerical convolution of the cumulative gain distribution can be performed on a computer which gives very good results even at very close ranges.

2. Near Field and Fresnel Zone Directive Antenna Coupling²

The electronic systems on naval ships are essential to modern warfare. Due to this fact there is a continued demand for more electronic systems, a large number of which have antennas associated with them. This creates a high potential for electromagnetic interference between systems, due to coupling between the various system antennas. This type coupling is particularly high between antennas in the same bands.

Some of the basic dimensions that must be analyzed are frequency, power, time, and space. The basic frequencies -- harmonics, spurious, and intermodulation -- can all cause considerable interference. The spatial dimension, i.e., relative separation and orientation, as well as scattering, has a considerable effect on the amount of interference. The absolute power of the radiating system is also of paramount importance.

The quantity which needs to be estimated is the coupling factor, which is a loss factor normally given in dB, and is determined from a function of several known parameters. The coupling factor (CF) is given by the following relationship:

CF (dB) =
$$10 \log \frac{P_t}{P_v}$$
,

where Pt is the transmitting power, and Pr is the received power.

The coupling factor can easily be misinterpreted since the effect of increased or decreased "coupling" values is the opposite of "coupling factor." An increase in coupling value corresponds to a decrease in coupling factor and, conversely, an increase in coupling factor corresponds to a decrease in coupling between the antennas.

When conditions are ideal, one can use the coupling factor equation to compute the coupling factor; however, in the practical situation the necessary conditions of far field, free space, and inband frequencies of operation seldom exist. All of these complications give rise to the statistical nature of the coupling factor and make it generally impossible to accurately obtain the coupling factor using a deterministic mathematical treatment. This has led to the use of continuous sampling techniques for determining the rean coupling factor for a given pair of antennas. A spectrum analyzer operating with the time base synchronized to the rotation of one radar is used. This displays the coupling level on a logarithmic scale for an azimuth setting and fixed test frequency of the second antenna. This display is photographed and mean level visually determined by placing a straight edge through the centers of the random variations. This gives a reasonably accurate coupling factor within ±1 dB. The test frequency or

the orientation of the fixed antenna is then changed to obtain other values. The average of the mean values then represents a mean coupling factor of a particular situation class for the specific antenna pair. In this manner a data base for mean coupling factor of antenna types and situations is generated.

B. Near Zone Radar Cross Section (RCS)3

The objective of this research was to define a generalized technique for ascernaining the near zone RCS of missiles. There are two basic problems which complicate near zone RCS determinations: (1) the target can be illuminated by both planar and nonplanar fields, and (2) the reflected fields can be nonuniform and nonplanar. The near zone RCS is a function of the following parameters: frequency, polarization, range, target size and aspect, and the radiation patterns of the receiver and transmitting antennas. A generalized formulation of the near zone is developed which shows that the near zone RCS is a function of the radiation patterns of the transmitting and receiving antennas, target geometry, and electrical properties which affect the currents induced in it. Sensitivity to the above parameters has been studied and indicates that each is important in varying degrees depending on the existing conditions.

Calculations of scattering properties for simple geometric shapes have been accomplished according to the Geometric Theory of Diffraction which give good agreement with near zone field measurements. Four techniques for determining characteristics of near zone RCS are presented: a direct measurement technique, a computational method, a "Subarea Matrix Method," and a "Modal Expansion Method." The first two can provide good cross section data under limited conditions; however, the third and fourth, in principle, have the capability of providing good data under generalized conditions.

Future work in the near zone RCS area will be concentrated on the Subarea Matrix Method or the Modal Expansion Method. The reason for this decision is that the direct measurement method is only good for specific encounters and the theoretical techniques are limited to simple geometric shapes.

Since neither of the recommended generalized techniques has been developed, it is not possible to guarantee that either will be totally successful. However, it is believed that the Modal Expansion Method has a good chance for success because it has been used successfully in determining radiation patterns of antennas. The data measurement techniques and equipment are presently available at Georgia Institute of Technology. In addition, some of the analytical and sampling criteria have already been established. None of these statements are true for the Subarea Matrix Method; however, the Subarea Matrix Method should not be completely discarded. Some simplified experiments would be worthwhile, although most of the effort is concentrated on the Modal Expansion Method. The method of direct measurement would also be of value for comparison in specific cases, along with the theoretical methods for simple geometric type targets.

C. Correction Factors for Near Field Gain Measurements4

The apparent gain of two antennas separated by a finite distance is different from the case of infinite separation, and many researchers have dealt with the problem of correcting for the effect. Most of these analytical methods make various assumptions about the fields which make it difficult to assign a value to the computational accuracy. This paper discusses the use of experimental data instead of assumed field distributions. The results are compared to the method used by Chu and Semplak. Both papers are based on the power transmission formula. The apparent gain of two antennas is given by

$$G_1 G_2 = \frac{(4\pi R)^2}{\lambda^2} \frac{P_r}{P_t}$$
,

where G_1 and G_2 are the apparent gains of antennas 1 and 2

R = distance between the antenna apertures

 λ = wavelength.

The equations used are exact; therefore, if the fields of the two antennas were known exactly on a common surface S and on the surfaces S_1 and S_2 which inclose antenna 1 and antenna 2, then the near field gain correction factors could be computed exactly.

In the method described in this paper the far field patterns of antennas 1 and 2 are the known data. Thus to obtain the fields of both antennas on a common surface, a spherical wave expansion is matched to the far field patterns. These expansions are then evaluated to determine the E and H field on the desired surfaces.

The approach of the authors was to use the patterns of one antenna transmitting with the other antenna removed. The problem with this approach is that multiple scattering is neglected. This is the only assumption made by this method. It has been shown that multiple scattering can have an appreciable effect. It is also common practice to make several measurements over a small range of antenna separation, and use curve fitting techniques to remove the scattering and multipath effects.

This approach has one other theoretical shortcoming. The method is sensitive to the truncation point of the spherical wave expansion. It is well known that spherical waves exhibit cutoff behavior that limits the number of modes to approximately N=ka, where k is the propagation constant, and a is the radius of the smallest sphere that can inclose the source. Two suggested guides for truncation are:

(1)
$$N = ka + 10$$

(2)
$$N = 7.7 (2a/\lambda)^{0.78}$$
 for $1 < a/\lambda < 15$.

In the first attempts the authors used N=25. It was later discovered that a better choice would have been between N=18 and N=22.

Accuracies on the order of ± 0.02 dB appear to be possible. The differences between computed and measured gain correction factors were a little higher, approximately 0.03 dB. This difference could be due in part to measurement error (see Figure 3, extracted from Reference 4). Agreement is quite good and the authors believe that future work will lead to even better agreement.

D. Near Field Measurements for Occupational Safety

1. Near Field Electric Energy Density Meter (EDM) 5

The EDM-2 electric energy density radio frequency survey monitor was designed to measure electric fields from 10 to 500 MHz. This NBS report describes the problems in collecting and interpreting the monitor readings. A number of sources of error are discussed:

- (1) The reactive near field components.
- (2) The amount of multipath interference.
- (3) Field polarizations.
- (4) The field modulation.
- (5) The complex interactions between the power source and nearby objects.

The relationship between the electric field and the magnetic field is completely ambiguous when measurements are conducted within one wavelength of the source.

This fact requires that both the electric and magnetic field be measured to obtain the total occupational health exposure. The EDM-2 monitor uses a set of three orthogonal dipoles to obtain the required isotropic response. The output of the dipoles is fed to the monitor electronics by special high resistance conductors to reduce the monitor's interaction with the RF field. The meter displays electric energy density from 0.003 to $30~\mu\text{J/m}^3$. These values equate to plane-wave equivalent power density from 0.1 to $1000~\text{mW/cm}^2$. The monitor is calibrated in units of microjoules per cubic meter. This choice was made because it simplified the circuit as well as providing useful RF data on an easily readable dial. The units are easily related to milliwatts per square centimeter of equivalent plane-wave field by the following relationships:

- (1) $S(\mu W/cm^2) = 60.0U_E(nJ/m^3)$
- (2) $S(mW/cm^2) = 60.0 U_E(\mu J/m^3)$
- (3) 10 mW/cm² (plane wave) = $0.3 \mu J/m^3$
- (4) $E(V/m) = 475.33 \left\{ U_E(\mu J/m^3) \right\}^{1/2}$
- (5) $E(V/m) = 15.03 \left\{ U_E(nJ/m^3) \right\}^{1/2}$

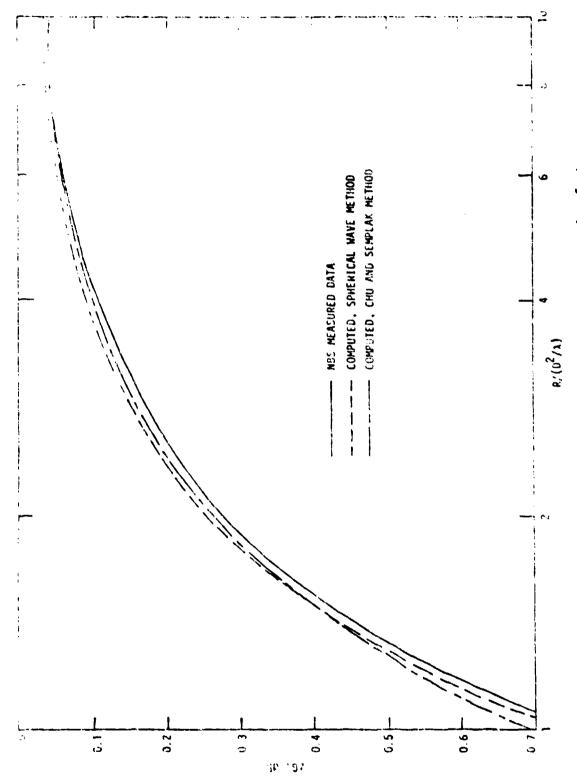


Figure 3. Measured and computed near field gain correction factors. (Extracted from Reference 4.)

The EDM-2 represents an improvement over existing monitors. It has a faster rise and fall time for pulse response, a dynamic range of 50 dB, reduced temperature sensitivity, low noise preamplifiers, and a response flatness of ±1 dB from 10 MHz to 500 MHz.

2. Measurement of Potentially Hazardous Electromagnetic Fields-- RF and Microwave⁶

The measurement of electromagnetic fields from 1 MHz to 100 GHz is discussed in this draft ANSI Standard, dated November 1978. This document was written to replace older documents which did not cover frequencies below 300 MHz or problems associated with the accurate measurements in the near field. This was due in part to the unavailability of instruments capable of making accurate measurements in the near field.

For the frequencies above 300 MHz meaningful data can be measured at distances greater than a^2/λ , where the probe dimension is less than a. a is the largest aperture dimension and λ is the wavelength.

Power density is used as the hazard indicator; however, no existing instrument measures power density directly. The techniques used are usually to measure one or more components of the electric field E or the magnetic field H and infer a power density based on the plane-wave far field relationships.

Measurements are required in the reactive or radiating near field region where standing waves occur. Since hazards are due to energy absorption, the parameters $|E|^2$ and $|H|^2$ can be measured. The energy densities $U_E = \epsilon^* |E|^2/4$ and $U_H = \mu^* |H|^2/4$ are also representative of the amount of hazard, where ϵ^* and μ^* are the real parts of the dielectric constant ϵ and the magnetic susceptibility μ , with the energy Joules/meter (J/m^3) . This choice conveniently allows both E and H to be expressed in the same units.

The typical situations in which near field measurements are required are:

- (1) Leakage fields from waveguides.
- (2) Radiation fields from large antennas.
- (3) Reactive fields from large low frequency horns.

"Leakage" refers to unintentional leakage of energy, whereas "radiation fields" refers to intentional radiations. The "reactive fields" exist in both the leakage and radiation fields and are generally stronger near inefficient radiators. The problem of multipath interference is another complication that exists to some degree in every situation.

Summary of Measurement Techniques. Some of the factors which determine the electromagnetic environment are: the direction of energy propagation with respect to the sources, polarization, frequency, type of modulation, and power of sources. The variable nature of these effects as

well as their effects on the resultant electromagnetic environment make it difficult to design and operate instruments which can measure the electromagnetic environment with sufficient accuracy to insule personnel safety.

The near fields of radio frequency sources are characterized by the combination of both reactive and radiation components. These components have spatial and temporal variations which are functions of the physical environment and the type sources that generate them. These variations create many unique situations which make calculations of near field intensities impractical. Therefore, one must rely on measurements.

III. COMMENTS ON NEAR FIELD LITERATURE SURVEY/CONCLUSIONS

In light of information contained in existing literature, it is obvious that the need discussed in Section I, i.e., electromagnetic interference in the near field of large microwave antennas, has not been investigated to sufficient depth to answer certain fundamental questions. Since these questions are of tremendous importance to a large number of defense systems and a growing number of commercial electronic systems, further research in this area seems appropriate. It is apparent that much of the material surveyed applies to near field testing phenomena. This material provides a sound basis and a strong impetus for future research.

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DEFINITIONS

- Antenna A device used to radiate or receive radio waves.
- Antenna Array An antenna system which uses multiple antennas to obtain directional effects.
- Antenna Gain, Relative The ratio of the power gain of an antenna relative to a half-wave dipole or isotropic antenna.
- Beamwidth, Half-Power The half power beamwidth is measured in the plane containing the lobe maximum. It is the angle between the two directions in that plane that intersects the 3 dB points of the maximum lobe.
- Dipole An antenna that produces a pattern that is similar to an elementary dipole. It is usually a $^1/_2$ λ straight radiator, which is fed at the center.
- Electric Field A vector field of the electric field strength.
- Electric Field Strength The magnitude of the electric field vecter.
- Far Field Region That part of the field of an antenna where the angular field distribution is nearly independent of the distance from the antenna, which is sometimes defined by $2D^2/\lambda$.
- Isotropic Having the same properties in all directions.
- Magnetic Field Strength The magnitude of the magnetic field vector.
- Magnetic Field Vector Given by the division of the magnetic induction by the permeability of the medium.
- Microwaves Normally are radio waves that range in frequency from approximately 1 GHz to 200 GHz.
- Near Field Region, Radiating The region of an antenna between the reactive near field region and the far field region. In this region the angular field distribution is dependent on the distances from the antenna. If an antenna is focused at infinity, this region is sometimes referred to as the Fresnel zone -- due to the analogy to optical terminology.
- Near Field Region, Reactive That part of the field immediately surrounding the antenna. For most antennas the region exists to a distance of $\lambda/2\pi$ from the antenna's surface, where λ is the wavelength.
- Radar Refers to a system which radiates electromagnetic waves and receives the reflections of the waves from distant objects. These reflections are utilized to determine the existence of the objects and their positions.

APPENDIX B

DESIGN, CONSTRUCTION, AND CALIBRATION OF PSEUDO-TRANSPARENT

NON-INTERFERING ELECTRIC AND MAGNETIC FIELD SENSORS

This entire appendix appeared as a paper in the Proceedings of IEEE Southeastcon 1981, pages 120-124 by Brian R. Strickland, George R. Edlin and Thomas H. Shumpert.

DESIGN, CONSTRUCTION, AND CALIBRATION OF PSEUDO-TRANSPARENT, NON-INTERFERING ELECTRIC AND MAGNETIC FIELD SENSORS

Abstract

For non-uniform, non-planar electromagnetic fields, knowledge of the magnitude and phase of each of the three orthogonal components of both the electric field and the magnetic field is required to completely characterize the field. Electrically small, polarization selective, electric and magnetic field sensors for measurement in the UHF band have been designed, constructed, and calibrated. These sensors are short dipoles (for electric fields) and small loops (for magnetic fields) with RF diodes across their gaps. The rectified RF is transferred down semiconductor (carbon-filled) lines to a DC voltmeter. Calibration of the sensors was accomplished in a "TEM cell" transmission line. Complete discussion of the various phases of this work is included in the paper.

Design

Two characteristics of the B Dot probe need to be considered: namely, frequency response and voltage output. (See Reference 3)

$$f_{\text{max}} = \frac{0.35c}{2\pi a} \tag{1}$$

where c is the speed of light and a is the radius of the loop, and

$$|v_{p}| = \frac{2\pi A E_{o} f}{c}$$
 (2)

where A is the area of the loop and E_0 is field density. Since

$$E_{O} = Bc \tag{3}$$

where B is the flux density in webers per square meter, equation (2) becomes

$$|V_{\mathbf{p}}| = 2\pi \text{ fAB} \tag{4}$$

Clearly, equations (1) and (4) are area dependent.

The voltage and frequency relationships of a dipole antenna can be derived from the equations given by Taggert and Workman (Reference 1) as follows:

$$\ell_{\text{eff}} = \frac{\lambda}{\pi} \tan \frac{\pi \hat{L}}{\lambda} \tag{5}$$

where $\ell_{\rm eff}$ = the effective length of the antenna, ℓ = the antenna half-length in meters. For a half-wave resonant antenna $\ell = \lambda/4$. For this condition, equation (5) reduces to

$$\ell_{\text{eff}} = \frac{\lambda}{\pi} \tag{6}$$

However, shorter non-resonant antennas cannot be treated in this manner as the effective length is reduced by the tangent function. For very short antennas using tan $0 \, ^{\circ} \, 0$ for small $0 \, l_{\rm eff} = l$ which indicates a region where $V_{\rm OC}$ is independent of frequency. This indeed was verified by the data. Furthermore,

$$V_{oc} = E^{\ell}_{eff} \tag{7}$$

where

 V_{oc} = open-circuit antenna voltage in volts

E = electric field in volts per meter.

Signal detection should occur at the probe as attempting to extract an RF signal from a field could introduce severe interference problems. It is suggested that the detecting element be mounted directly across the gaps in the case of the B Dot probe and directly between the elements of the dipole antenna (see Figure 1).

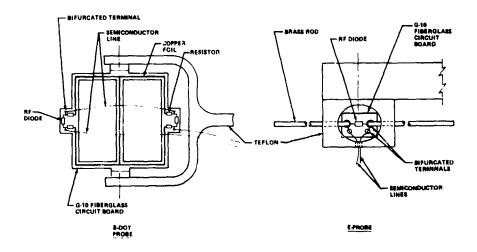


Figure 1. Geometrical and electrical configuration of E and B field probes.

Clearly, no metallic conductor should be used in removing the detected signal from the field. However, carbon filled, or semiconductor lines, may be used with virtually no effect on the field. One disadvantage of semiconductor lines is the high resistance/length. Typical values of resistance for 5-10 meters is 300 K Ω . This problem can be alleviated by either use of a very high impedance voltmeter (we used 100M Ω) or use of precision high input impedance buffer amplifiers.

The above equations, 1 through 7, were used to obtain dimensions of the two probes for operation in the desired frequency range of interest. Another factor to consider is the requirement for noninteracting or nonperturbing probes. This placed an upper limit on the dimensions.

Construction

Construction of a ruggedized B Dot as well as E probes was accomplished. Teflon was used for all possible portions of the probe. The elements of the probes are constructed from single sided G-10 epoxy fiberglass printed circuit board. The B dot probe used the double gap design to cancel the effect of the E field across the gap. The major change in design was to go to a two dimensional design since transmission line matching was no longer required. Also because of the ease of construction, a square instead of circular loop was used. Therefore, equation 1 will be modified slightly for the new probes. The E field probe used a 3/32 brass rod for the elements. For each probe, the dimensions were kept short ($\ell < .1\lambda$) to assure that the probe does not resonate. IN832 RF diodes were used in both cases: two for the B dot and one for the E field probe.

Accurate measurement of the orthogonal components of the B and E fields requires that the probes are capable of horizontal, radial or vertical orientation with the same phase center. Therefore, freedom of movement is necessary in three planes and the probe mounts were constructed to provide this movement without changing the phase center. Figure 1 gives the details of the probes.

Calibration

There are several methods of calibration of field measuring probes (Reference 1). Among these are the standard antenna method, the standard field method and the injection method. Basically, the standard field technique involves placing the probe in a known standard field and determining a calibration factor or antenna coefficient from the magnitude of the known field and the output of the probe. The antenna equations must be solved for 'E', and

$$K = \frac{E_{m}}{E_{3}}$$
 (8)

where

K = the antenna coefficient

 $E_m =$ the measured field

 $E_k =$ the known field.

To calibrate an antenna using the standard antenna method, the calibrating field is measured first using a standard antenna. The antenna under test is then substituted in place of the standard antenna. From these data, the antenna coefficient can be determined. With the injection method, a low

impedance voltage source (less than 0.1 ohm) is used to inject a known voltage in series with the antenna. By calculating the effective length of the antenna and knowing the injected voltage, the antenna coefficient can be determined. This method is the least accurate, while the standard field method is the most accurate. A complete discussion of calibration techniques can be found in reference (1).

Calibration was performed using the standard field method. A Narda Microwave Corporation TEM Transmission Cell was used to establish the standard field. With this TEM cell the standard field method can be used with great accuracy since the only field perturbations are caused by the probe itself and no other sources of reflection and interference such as the ground are present.

TEM Cells (Reference 2) are sections of a two conductor transmission line which operate in the transverse electromagnetic mode. TEM cells are sometimes referred to as Crawford Cells after their developer, Myron Crawford of the National Bureau of Standards. The cell is structured from a rectangular outer conductor and a flat center conductor. The center conductor, or septum, is located midway between the top and bottom walls. The ends of the cell taper down to precision type N connectors at each end. Dimensions and tapers of the cell are chosen to provide a 50 ohm impedance from end to end. The field at the center of the cell is uniform and very close to the 377 ohm impedance of free space. The electric field intensity in volts per meter at the center of the cell, midway between the septum and outer wall is given by the equation

$$E = 47.1 \sqrt{P_n} \tag{9}$$

where P_n is the net power transmitted thru the cell. The electric field is polarized normal to the plane of the septum. The probe to be calibrated is placed in the cell and the output voltage is measured thru the ports provided in the cell.

It was desired to obtain a quick and accurate calibration technique. The first step in this procedure was to obtain a general idea of probe output voltage variations with frequency, field level, temperature, and RF diodes. This was accomplished by hand measurement of Vo while the four parameters were varied. The TEM cell was placed inside an environmental chamber where temperature and humidity could be varied and measured. The humidity was kept relatively constant. Five E field probes and five B field probes were used to obtain the effect of different RF diodes on Vo. If Vo were plotted vs. frequency, field level, and temp., we would have a three parameter family of curves which would require a three dimensional graph to display. Since the desired end result was to obtain

$$F_{\tau} = f(f, V_0, T) \tag{10}$$

where

f = frequency

 F_{τ} = field level (B or E)

Vo = Probe output voltage

T = temperature,

we needed to determine the interdependence of variables and determine a way to computer process the data to obtain equation (10). A standard polynomial regression technique would not work because it does not extend to more than one variable. Also it was not clear how to present the data in two dimensional graphs to best show off the data and bring out significant hidden facts and data trends. As a result many different presentations were examined, i.e., Vo vs. f with F_{L} as a parameter, F_{L} vs. Vo with f as a parameter, F_{L} vs. Vo with T as a parameter, etc. Since we have four different parameters, 4! = 24 number of possible permutations exist, i.e., there are 24 possible ways to present the data.

After examination of the data the temperature variations seemed to vary from diode to diode and no pattern could be recognized. The temperature variation seemed to be a random function of the diode. All 24 permutations were not used (8 would have T as a variable) so it is possible that a pattern could exist and merely not be discovered. This aspect will be investigated more fully at a later date. Based on this it was determined to leave T out and determine equation (10) with T fixed. Either equations could be determined at several different temperatures or an equation could be determined immediately prior to use at ambient temperature. This is feasible since determination of the equation will only take about 10-30 minutes. Styrofoam would need to surround the probes for ruggedization and to prevent heating or cooling due to wind or direct sunlight. Removal of T as a variable greatly reduces the data problem. Now only 6 different permutations exist (actually only 3). It was decided to plot Vo vs. f with FL as a parameter. For brevity only the E field probe is discussed in the remainder of the paper. A typical plot is shown in Figure 2.

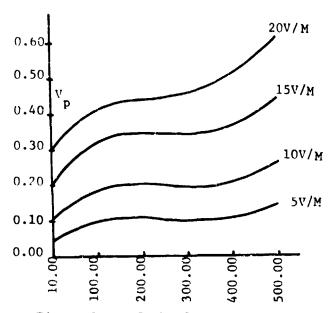


Figure 2. E Probe frequency response.

The Hewlett-Packard 9825 microcomputer is utilized to control the operation and the Hewlett-Packard Interface Bus is used throughout the system. The block diagram of the computer calibration system is shown in Figure 3. Data is taken by the computer as follows:

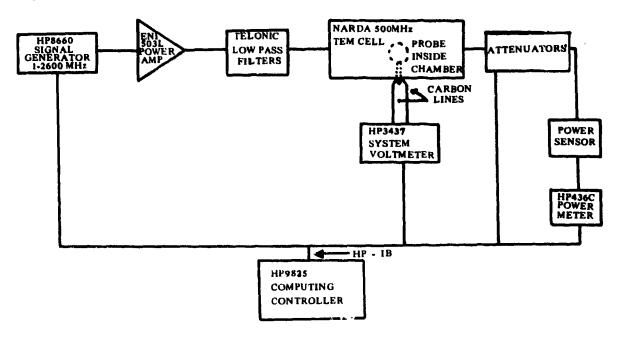


Figure 3. Probe calibration configuration.

- 1. A given F_L is set within $\pm .5 \ db$ (since the signal generator has a ldb step attenuator).
 - 2. The frequency is set.
 - 3. Vo, f and FL are stored in an array.
 - 4. The frequency is stepped and 3 repeated.

Assume for illustration purposes 9 frequency points and 9 field levels are used. This gives a 9 x 9 x 3 tensor or 3 dimensional array with 81 rows. If f and F_L are independent, we can change equation (10) to

$$F_{L} = f(f) f (Vo) . \tag{11}$$

This is approximately true and will lend itself to polynomial regressions around a given F_L and f, and if f and F_L are not allowed much deviation from these values, equation (11) will give negligible error. However, a more general and accurate equation was desired so we must go back to equation (10).

Since f is the only truly accurate variable left to us (F_L is \pm .5 db from the set value sc), a polynomial regression is performed on the data 9 times for each frequency point giving a series of equations as follows (assuming 3rd order polynomial regression):

$$F_{L} = c_{11} vo^{3} + c_{12} vo^{2} + c_{13} vo , f = f_{1}$$

$$F_{L} = c_{21} vo^{3} + c_{22} vo^{2} + c_{23} vo , f = f_{2}$$

$$F_{L} = c_{91} vo^{3} + c_{92} vo + c_{93} vo , f = f_{9} .$$
(12)

Now the original 9 x 9 x 3 tensor is discarded if program memory is a problem and we have a new 9 x 4 matrix composed of the C's in equation (12). Note in equation (12) that a constant term does not exist. This is because it is desired to force $\mathbf{F}_{L} = \mathbf{0}$ when $\mathbf{Vo} = \mathbf{0}$. Now a single equation is obtained from equation (3) by using a polynomial regression on each column of the new 9 x 9 matrix giving 3 equations for the 3 columns as a function of frequency

$$F_L = C_1(f)Vo^3 + C_2(f)Vo^2 + C_3(f)Vo$$
 (13)

Referring to equation (13) and Figure 2, it is expected that F_L will follow a -f curve, therefore $C_1(f)$, $C_2(f)$, $C_3(f)$ will follow an f^3 curve which indicates that $C_1(f)$, $C_2(f)$, $C_3(f)$ need to be 3rd degree polynomials. It is also noted that F_L vs. Vo is also linear as is expected from equation (7) so the first two columns of the matrix in equation (12) can be dropped without causing appreciable error if computer memory or time is a constraint.

This type of analysis is also used with the B dot probe to determine the degrees of the four polynomials used.

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APPENDIX C

THE DESIGN AND CONSTRUCTION OF A PROBE

POSITIONER FOR ELECTROMAGNETIC FIELD MEASUREMENTS

Report of MICOM Technical Report RT-80-18, June 1980, George R. Edlin and Richard Aldridge.

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I. INTRODUCTION

The purpose of the project is to design and construct a test fixture to measure the characteristics of the near field emitted from a standard gain horn. This report consists of a discussion of the probes used for measurement, possible designs of the support structure, methods of positioning the probes, and data acquisition techniques. A discussion of a working model is also included.

II. DISCUSSION OF ELECTRIC AND MAGNETIC PROBES

Characterization of an electromagnetic field requires that the orthogonal components of both the B and E fields be accurately measured and recorded. This is true for the far field as well as the near field environment. These measurements can then be used to determine the power density of the field. Conventional measurement of these fields is performed using the B Dot probe (antenna) and calibrated dipole antennas (E probes).

Two characteristics of the B Dot probe need to be considered; namely, frequency response and voltage output. By the equations given in Reference 1

$$f \max = \frac{0.35 \text{ c}}{2\pi \text{ a}} \tag{1}$$

where c is the speed of light and a is the radius of the loop, and

$$|Vp| = \frac{2\pi AE_0f}{c}$$
 (2)

where A is the area of the loop and $E_{\rm O}$ is the field density. Since

$$E_{O} = Bc \tag{3}$$

where B is the flux density in webers per square meter, equation (2) becomes

$$|VP| = 2\pi fAB \tag{4}$$

clearly, equations (1) and (4) are area dependent.

The voltage and frequency relationships of a dipole antenna can be derived from the equations given by Taggert and Workman (Reference 2) as follows:

$$1_{\text{eff}} = \frac{\lambda}{\pi} \tan \frac{\pi \ell}{\lambda} \tag{5}$$

where

 1_{eff} = the effective length of the antenna,

the antenna half-length $(\frac{\lambda}{4})$ in meters for a self-resonant antenna (half wavelength).

For this condition, equation (5) reduces to

$$1_{eff} = \frac{\lambda}{\pi} .$$

However, shorter nonresonant antennas cannot be treated in this manner as the effective length is reduced by the tangent function. Furthermore,

$$V_{oc} = El_{eff}$$
 (6)

where

V = open-circuit antenna voltage in volts,

E = electric field in volts per meter,

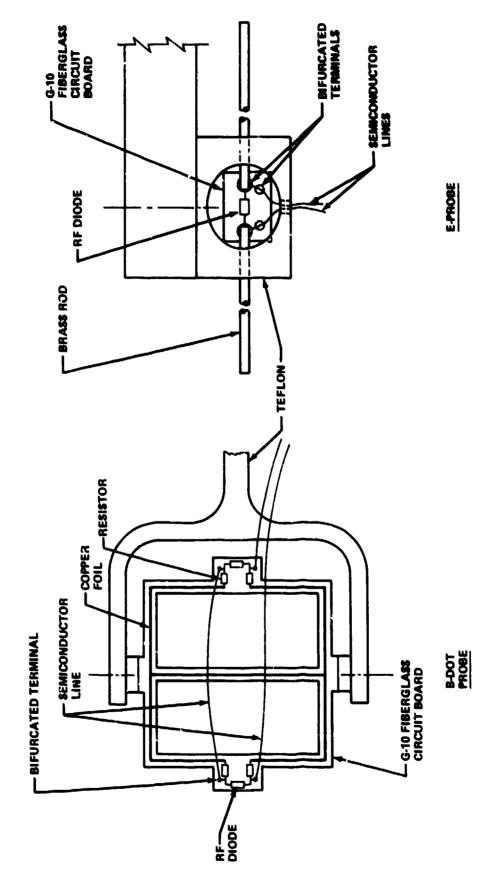
 l_{eff} = effective antenna length in meters.

Accurate measurement of the orthogonal components of the B and E fields requires that the probes are capable of horizontal, radial or vertical orientation with the same phase center. Further, the probes must be moved spatially throughout the field; that is, in the longitudinal and transverse planes. This introduces physical mounting problems which will be considered later.

Signal detection should occur at the probe as attempting to extract an RF signal from a field could introduce severe interference problems (Reference 3). It is suggested that the detecting element be mounted directly across the gaps in the case of the B Dot probe and directly between the elements of the dipole antenna (see Figure 1). RF diodes, light emitting diodes or liquid crystals could be successfully used for detection. A filter may be desirable, but is not essential for accurate field measurements.

Clearly, no metallic conductor should be used in removing the detected signal from the field. However, carbon filled or semiconductor lines, may be used with virtually no effect on the field. One disadvantage of semiconductor lines is the high loss introduced into the system. While not excessive, this loss must be considered in the probe calibration factor. In the case of the light-emitting diode or liquid crystal detectors, fiber optic lines could be used to transmit the signal out of the field with the advantage of a high degree of isolation between the probe and the load of the measuring device. This method, however, is much more complex and the cost is considerable. If cost and complexity can be overlooked, the fiber optic system seems to be the most desirable. Another possible configuration would be to convert the detected signal to a frequency modulated audio tone to be transmitted out of the field. No effort to prove or disprove this method was expended during this research, but it certainly seems to warrant future investigation.

There are several methods of calibration of field measuring probes (Reference 2). Among these are the standard antenna method, the standard field method and the injection method. Basically, the standard field



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Figure 1. Geometrical and electrical configuration of E and B field probes.

technique involves placing the probe in a known standard field and determining a calibration factor or antenna coefficient from the magnitude of the known field and the output of the probe. The antenna equations must be solved for 'E', and

$$K = \frac{E_{m}}{E_{k}} \tag{7}$$

where

K = the antenna coefficient

E = the measured field

 E_{ν} = the known field.

To calibrate an antenna using the standard antenna method, the calibrating field is measured first using a standard antenna. The antenna under test is then substituted for the standard antenna. From these data, the antenna coefficient can be determined. With the injection method, a low impedance voltage source (less than 0.1-ohm) is used to inject a known voltage in series with the antenna. By calculating the effective length of the antenna and knowing the injected voltage, the antenna coefficient can be determined. This method is the least accurate, while the standard field method is the most accurate. A complete discussion of calibration techniques can be found in Reference 2.

III. DESIGN OF PROBE SUPPORT STRUCTURE

The primary problem encountered in measuring an electromagnetic field is distortion of the field by the measuring equipment. Metallic materials are completely ruled out for use as a probe support structure since ground planes are easily established on metallic surfaces, particularly at higher frequencies. Therefore, it is desirable to use materials which are insulators.

Several possibilities exist for probe support using non-metallic materials such as ceramics, wood, and many types of plastics. While ceramic material is useful, cost and fragility would limit its use in ruggedized systems. Wood has been in use for such applications for a considerable period of time; however, there are some disadvantages. Dimensional instability due to moisture, heat, and light make wood a difficult material to use for probe positioners. Plastic while far from ideal, seems to be one of the better alternatives. Unless extremely heavy grades of plastic are used, the slightest wind can cause it to flex. If plastic materials have an advantage, it is their ability to endure exposure to weather.

IV. PROBE POSITIONING AND DATA ACQUISITION

Another major problem encountered in field characterization is position control and data acquisition. For the present project, the required precision is \pm five millimeters. Measurements will be made in three directions and in three orientations. Over the required distances, this amounts to some 216,000 data points. The mass of data and the required precision

dictate that a computer be used to control the test. A complete discussion of computer control is beyond the scope of this paper; however, a summary of the method used will be discussed later.

Position indication can be accomplished by analog or digital means. Analog methods involve the use of a multi-turn potentiometer in the motor drive train. Several problems are encountered with this method. Slippage, motor starting torque and the necessity of A to D conversion make this method undesirable.

By far, the better method of position indication is direct digital pulse generation. Pulses can be generated either optically, electrically, or mechanically. The optical method involves a light source passing over a coded area of the positioner frame such that the light might be reflected or not reflected from the frame. A photo diode could be used to receive the light pulses. These light pulses could also be transmitted via a fiber optic line to the controlling system. Encoded electrical tones could be used to indicate position by much the same methods. However, for both optical and electrical methods, decoding at the computer introduces an additional processing step. A mechanical position indicator could also be used. One method is to use a small micro-switch with a roller arm that rolls over trip points on the stationary portion of the positioner frame. These switch closures are simply counted by the computer so that the computer can keep track of where the positioner is located. Again, the use of semiconductor lines is necessary to avoid disturbance of the electromagnetic field being measured.

Construction of the probe positioner began with the stationary frame. Six-inch poly vinyl chloride (PVC) pipe was used for the support uprights and the rails. Nylon bolts and PVC cement were used to assemble the pipe. Since PVC pipe will sag unless supported, several braces were installed using metal screws for rigidity. Other fastening methods may have been more desirable from an interference standpoint, but these small amounts of metal close to the ground level did not adversely affect the field.

The dimensions of the frame are shown in Figure 2. The length of the frame was conveniently chosen to make use of standard lengths of PVC. This length allows a measurement distance of 4.8 meters which is adequate for the test.

It was necessary to be able to move the positioner periodically during construction and testing. However, because of the positioner's weight, swivel casters were installed to facilitate movement. To construct these casters, a round wooden block was placed upright inside the support and a plastic plate attached to the bottom of the block. The wheel assembly was then attached to the plastic plate.

The mast trolly assembly was constructed of PVC and two-inch redwood and was assembled entirely with nylon hardware (Figure 3). The runners were made by cutting a 220° segment of eight-inch PVC pipe. The redwood was used for crossmembers. The mast bracket is 3/8" PVC sheet with redwood spacers. Wheel mounting brackets were cut from plexiglass blocks. When using these materials, careful and accurate alignment was absolutely essential.

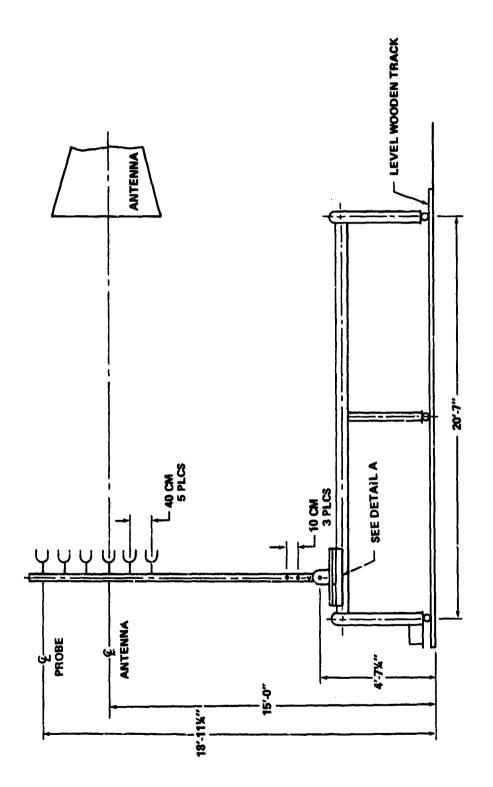


Figure 2. Dimensional sketch of probe positioner.

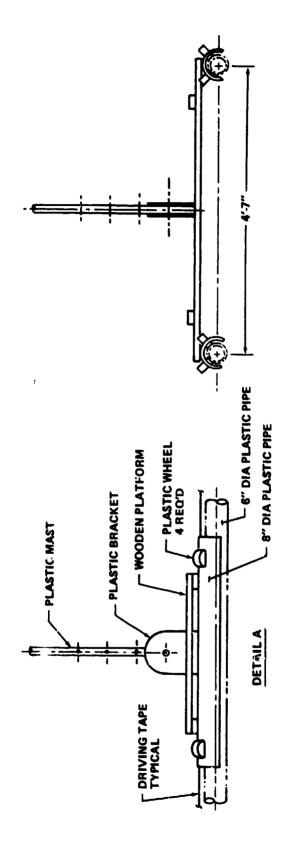


Figure 3. Detail of mast trolley.

The mast consisted of a piece of two-inch PVC pipe. Guy ropes were necessary to keep the mast from bending as the trolly moved along the positioner.

According to Reference 5, "Use of B Dot as a standard field measuring probe has given rise to a need for a ruggedized probe. The development of a more durable probe would save considerable time on repair and checkout of damaged probes." With this in mind, construction of a ruggedized B Dot as well as E probes was accomplished. Teflon was used for all possible portions of the probe. The element of the B Dot probe is constructed from single sided G-10 epoxy fiberglass printed circuit board. The output of this probe compares favorably with the output of a standard B Dot probe of comparable size. A similar construction technique was used for the E probe. A 3/32" brass rod was used for the elements. For each probe, the length is kept short ($\ell < 0.1\lambda$) to ensure that the probe does not resonate. Freedom of movement is necessary in three planes and the probe mounts are constructed to provide this movement without changing the phase center. Figures 1, 4, 5, 6 and 7 give the details of the probes which are mounted to the mast with 1/2" nylon bolts. All semiconductor lines from the probes are brought to a common junction box at the base of the mast where they are converted to shielded lines. Two separate masts were constructed to reduce probe damage during testing. The B Dot probes are mounted on one mast and the E probes are mounted on the other.

At the time of this writing, calibration of the probes had not been completed. However, an outline of the procedure to be used follows.

Calibration will be performed using the standard field method. A Narda Microwave Corporation TEM Transmission Cell will be used to establish the standard field. TEM Cells (Reference 4) are sections of a two conductor transmission line which operate in the transverse electromagnetic mode. TEM Cells are sometimes referred to as Crawford Cells after their developer, Myron Crawford of the National Bureau of Standards. The cell is structured with a rectangular outer conductor and a flat center conductor. The center conductor, or septum, is located midway between the top and bottom walls. The ends of the cell taper down to precision type N connectors at each end. Dimensions and tapers of the cell are chosen to provide a 50-ohm impedance from end to end. The field at the center of the cell is uniform and very close to the 377-ohm impedance of free space. The electric field intensity in volts per meter at the center of the cell, midway between the septum and outer wall is given by the equation

$$E = 47.1 \sqrt{P_n}$$
 (8)

where $P_{\rm n}$ is the net power transmitted thru the cell. The electric field is polarized normal to the plane of the septum. The probe to be calibrated is placed in the cell and the output voltage is measured through the ports provided in the cell.

A motor assembly is installed at the base of one of the support uprights of the positioner frame using a gear reduction and tape drive system to move the mast trolley. The tape used is plastic banding material which does not stretch appreciably and is very durable. It was determined that moving the

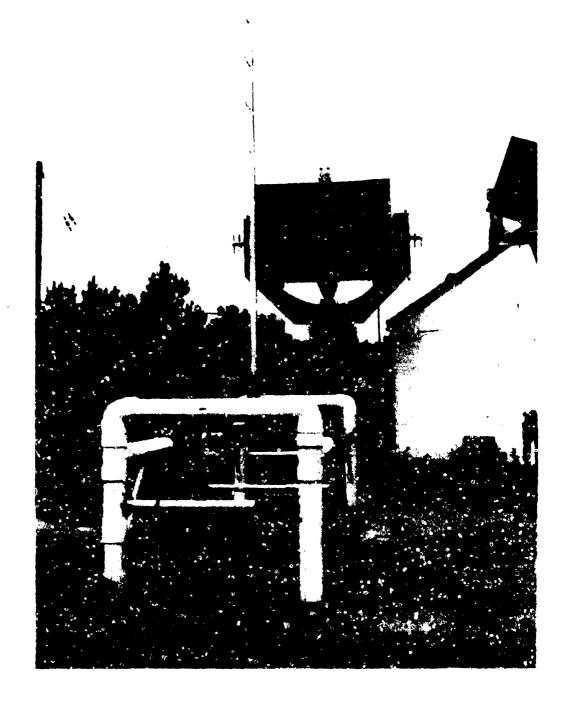


Figure 4. Actual probe positioner in foreground with UHF high gain pyramidal horn in background.

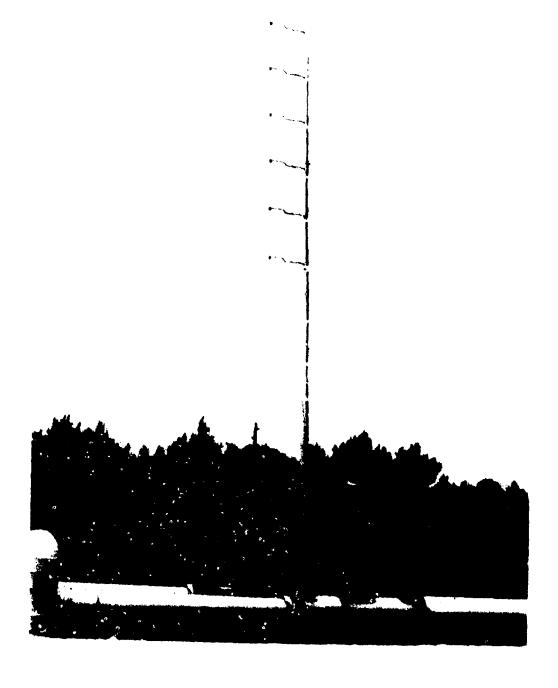


Figure 5. Closeup of mast and trollev assembly.

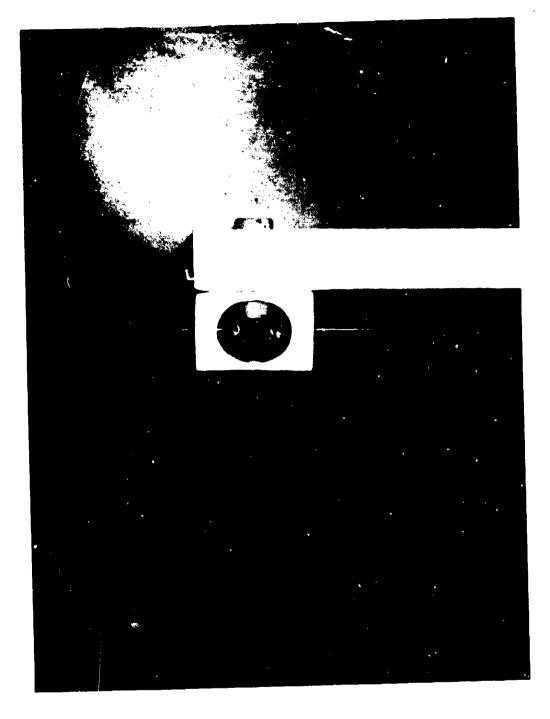


Figure 6. The E-probe and support bracket with carbon-filled (semiconductor) signal lines.



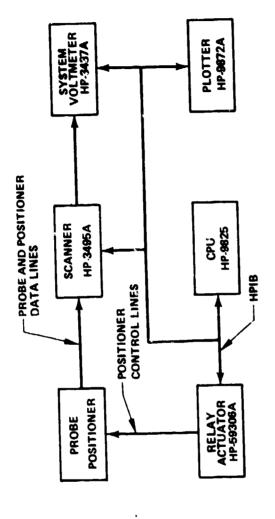
Figure 7. The B dot probe and support bracket.

trolley at a slow constant speed rather than stopping and starting gave much more accurate measurements. This was due to the springiness of the mast. With the system used, the trolley moves at a speed of 8.48 centimeters-per-second. Position is indicated to the computer using the microswitch system described previously. There are six probes on the mast which makes a total of seven voltages that the computer must recognize and measure for each data point. Data points are every ten centimeters, or approximately one data point every 1.8 seconds. Since the computer requires a very small fraction of this time to make the required measurements, the amount of error is nearly negligible. The error that does exist can be accounted for in the data uncertainties.

Figure 8 is a system block diagram of the test setup. The Hewlett-Packard 9825 Microcomputer is utilized to control the operation and the Hewlett-Packard Interface Bus is used throughout the system. Complete analysis of the computer system would serve no purpose here as it is the subject of another report, as is the resulting data.

Design and construction of a low interference probe positioner is feasible. There is considerable demand for such devices in the field of electromagnetic research and several proposals for its utilization already exist.

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Figure 8. System block diagram.

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AMPENDIX D

ANALYTICAL DERIVATION OF THE APERTURE DISTRIBUTION MODEL

APPENDIX D

ANALYTICAL DERIVATION OF THE APERTURE DISTRIBUTION MODEL

The E-field aperture distribution used for the theoretical model is the waveguide TE_{01} , i.e., the E-field is described using a quadratic phase term and a cosine amplitude term. This aperture distribution is used with vector Smythe-Kirchhoff formula for diffraction to develop the spatial representation of the radiated E-field. The vector Smythe-Kirchhoff is given in equation (D-1),

$$\dot{\vec{E}}(x,y,z) = \frac{1}{2\pi} \nabla x \int (\hat{n} X \bar{E}) \frac{e^{jkR}}{R} da' . \qquad (D-1)$$
Aper

Figure 1 shows the horn geometry and aperture associated coordinate system.

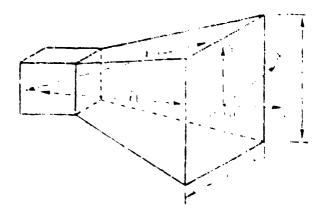


Figure 1. Geometry and Coordinate System of UNF Pyramidal Horn.

The TE_{01} Waveguide E-field distribution is given by equation (D-2).

$$\vec{E}(x,y,0) = E_0 \cos \frac{\pi x}{a} \exp \left[-jk \left(\frac{x^2}{2\ell_H} + \frac{y^2}{2\ell_E} \right) \quad \hat{U}_y \right]. \quad (D-2)$$

Let $\hat{n} = \hat{U}_z$ then:

$$\hat{\mathbf{n}} \mathbf{X} \stackrel{\triangleright}{\mathbf{E}} = -\hat{\mathbf{U}}_{\mathbf{X}} \mathbf{E}_{\mathbf{0}} \cos \frac{\pi_{\mathbf{X}}}{\mathbf{a}} \exp \left[-\mathbf{j} \mathbf{k} \left(\frac{\mathbf{x}^2}{2 \hat{\mathbf{l}}_{\mathbf{H}}} + \frac{\mathbf{y}^2}{2 \hat{\mathbf{l}}_{\mathbf{E}}} \right) \right] .$$

The relationship of R, r, r', and z are shown in Figure 2.

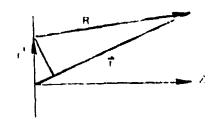


Figure 2. Source - observation coordinate system.

Where:

$$R = |\vec{r} - \vec{r}|$$
, $R = \left[z^2 + (x - x^{\dagger})^2 + (y - y^{\dagger})^2\right]^{1/2}$. (D-3)

Where (x,y,z) is the observation point and (x',y',o) is the source point, then we have

$$\frac{a}{2} \frac{b}{2}$$

$$\vec{E} (x,y,z) = -\frac{E_0}{2\pi} \nabla x \int \int U_x \cos \frac{\pi x'}{a}$$

$$-\frac{a}{2} - \frac{b}{2}$$

$$\exp \frac{\left[-jk\left(\frac{x^{12}}{2^{\frac{n}{k}}} + \frac{y^{12}}{2^{\frac{n}{k}}} - \left(z^{2} + (x-x^{1})^{2} + (y-y^{1})^{2}\right)^{1/2}\right)\right]}{(z^{2} + (x-x^{1})^{2} + (y-y^{1})^{2})^{1/2}} dx^{1}dy^{1} . \quad (D-4)$$

Let
$$\vec{A} = \hat{U}_{x} \cos \frac{\pi x'}{a} \exp \left[-jk \left(\frac{x'^{2}}{2 \hat{k}_{H}} + \frac{y'^{2}}{2 \hat{k}_{E}} - (z^{2} + (x-x')^{2} + (y-y')^{2})^{1/2} \right) \right]$$

$$\left[z^{2} + (x-x')^{2} + (y-y')^{2}\right]^{1/2} . \tag{D-5}$$

If
$$K = \cos \frac{\pi x'}{a} \exp \left[-jk \left(\frac{\chi'^2}{2\ell_H} + \frac{y'^2}{2\ell_E} \right) \right]$$
 (D-6)

then substituting K in equation (D-5) yields:

$$\hat{A} = \hat{U}_{x} \frac{K \exp \left[-jk(z^{2} + (x-x')^{2} + (y-y')^{2})^{1/2}\right]}{\left[z^{2} + (x-x')^{2} + (y-y')^{2}\right]^{1/2}}.$$
(D-7)

Therefore:

$$\nabla \mathbf{X} \hat{\mathbf{A}} = \frac{\partial \mathbf{A}_{\mathbf{X}}}{\partial \mathbf{z}} \hat{\mathbf{U}}_{\mathbf{y}} - \frac{\partial \mathbf{A}_{\mathbf{X}}}{\partial \mathbf{y}} \hat{\mathbf{U}}_{\mathbf{z}}$$
 (D-8)

and

$$\frac{\partial A}{\partial z} = K \frac{jkz}{R^2} - \frac{Kze^{jkR}}{R^3}$$

and

$$\frac{\partial A_{x}}{\partial y} = \frac{Ke^{jkR}jk(y-y')}{R^{2}} - \frac{K(y-y')e^{jkR}}{R^{3}} . \tag{D-10}$$

Thus

$$\nabla x \bar{A} = K \left[\hat{U}_{y} \left(\frac{jkze^{jkR}}{R^{2}} - \frac{ze^{jkR}}{R^{3}} \right) + U_{z} \left(\frac{K(y-y')e^{jkR}}{R^{3}} - \frac{jk(y-y')e^{jkR}}{R^{2}} \right) \right]$$
(D-11)

The curl is taken inside the integral by the application of the Leibnitz rule.

$$\vec{E}(x,y,z) = -\frac{E_0}{2} \iint \nabla x \vec{A} dx' dy' . \qquad (D-12)$$

Substituting $\nabla X \vec{A}$ into equation (D-12) and taking the negative sign inside yields:

$$\stackrel{\rightarrow}{E}(x,y,z) = \frac{E_0}{2\pi} \int_{X'=-a/2}^{a/2} \int_{X'}^{b/2} \left(\frac{ze^{jkR}}{R^3} - \frac{jkze^{jkR}}{R^2} \right) + \frac{ikze^{jkR}}{R^2} + \frac{ikze^{jkR}}{R^2} + \frac{ikze^{jkR}}{R^2} \right) + \frac{ikze^{jkR}}{R^2} + \frac{i$$

$$\hat{\mathbf{u}}_{\mathbf{z}} \left(\frac{\mathbf{j} \mathbf{k} (\mathbf{y} - \mathbf{y}') \mathbf{e}^{\mathbf{j} \mathbf{k} \mathbf{R}}}{\mathbf{R}^2} - \frac{(\mathbf{y} - \mathbf{y}') \mathbf{e}^{\mathbf{j} \mathbf{k} \mathbf{R}}}{\mathbf{R}^3} \right) \right] \quad d\mathbf{x}' d\mathbf{y}'$$
 (D-13)

where

$$R = \left[z^2 + (x - x')^2 + (y - y')^2\right]^{1/2}, \qquad (D-14)$$

$$R' = jkRy - y - jkRy' + y' = (y - y') + j(kRy - kRy')$$
, (D-15)

$$R_2 = z - jkRz ag{D-16}$$

$$K' = Ke^{jkR}$$
(D-17)

$$K' = \cos \frac{\pi x'}{a} \left\{ \cos \left[k \left(\frac{x'^2}{2 \ell_H} + \frac{y'^2}{2 \ell_E} - R \right) \right] - j \sin \left[\left(\frac{x'^2}{2 \ell_H} + \frac{y'^2}{2 \ell_E} - R \right) \right] \right\}$$
(D-18)

$$B = k \left(\frac{x^{\prime 2}}{2 \ell_H} + \frac{y^{\prime 2}}{2 \ell_E} - R \right)$$
 (D-19)

$$\frac{d}{E}(x,y,z) = \frac{E_{o}}{2\pi} \hat{U}_{y} \int_{x'=-a/2}^{a/2} \int_{x'=-b/2}^{b/2} \frac{K'R_{2}}{R^{3}} dx'dy' + \frac{1}{2} \int_{x'=-a/2}^{a/2} y'=-b/2$$

$$\frac{E_{0}}{2\pi} \hat{U}_{z} \int_{x'=-a/2}^{a/2} \int_{y'=-b/2}^{b/2} \frac{K'R'}{R^{3}} dx'dy'$$
(D-20)

$$K'R_2 = z \cos \frac{\pi_X'}{a} \left\{ [\cos B - kRsinB] - j [\sin B + kRcos B] \right\}$$
 (D-21)

$$K' = \cos \frac{\pi x'}{a} (\cos B - j \sin B)$$
 (D-22)

$$K'R' = \cos \frac{\pi_{X'}}{a} \left\{ [(\cos B) (y' - y) + (\sin B) (kRy - kRy')] + \right.$$

$$j [(kRy - kRy') (\cos B) - (y' - y) (\sin B)] \right\}. \tag{D-23}$$

Let

$$B_1 = \cos \frac{\pi x'}{a} \cos B \tag{D-24}$$

$$B_2 = \cos \frac{\pi x'}{a} \sin B \tag{D-25}$$

$$B_3 = R^3 \tag{D-26}$$

$$B_{4} = (y - y') kR$$
 (D-27)

Substituting these expressions into equation (D-20) produces

$$\bar{E}(x,y,z) = \frac{E_0}{2\pi} \hat{U}_y \int_{x'=-a/2}^{a/2} \int_{y'=z-b/2}^{b/a} \frac{z}{B_3} \left(B_1 - kRB_2\right) dx'dy'$$

$$- j \int_{x'=-a/2}^{a/2} \int_{y'=b/2}^{b/2} \frac{z}{B_3} \left(B_2 + kRB_1 \right) dx'dy'$$

$$+ \frac{E_{o}}{2\pi} \hat{U}_{z} \begin{bmatrix} \frac{a/2}{\int_{x'=-a/2}^{b/2} \int_{y'=-b/2}^{b/2} \left(\frac{y-y'}{B_{3}}\right) \left[-B_{1} + kRB_{2}\right] dx'dy'}$$

$$+ j \int_{x'=-a/2}^{a/2} \int_{y'=-b/2}^{b/2} \left(\frac{y-y'}{B_3}\right) \left[B_2 + kRB_1\right] dx'dy'$$
 (D-28)

For further simplification the following functions are defined:

$$F_1 = \frac{z}{2\pi B_3} \left(B_1 - kRB_2 \right)$$
 (D-29)

$$F_2 = -\frac{z}{2\pi B_3} \left(B_2 + kRB_1 \right)$$
 (D-30)

$$F_3 = \left(\frac{y - y'}{2\pi B_3}\right) \left[-B_1 + kRB_2\right] \tag{D-31}$$

$$F_{4} = \left(\frac{y - y'}{2^{\pi}B_{3}}\right) \left[B_{2} + kRB_{1}\right] . \qquad (D-32)$$

Substituting the above functions into equation (D-28), one obtains

$$\vec{E}(x,y,z) = \vec{E}_0 \left\{ \hat{U}_y \left[\iint_{Aper} F_1 \, dx' dy' + j \iint_{Aper} F_2 \, dx' dy' \right] + \hat{U}_z \left[\iint_{Aper} F_3 \, dx' dy' + j \iint_{Aper} F_4 \, dx' dy' \right] \right\}.$$
(D-33)

Using the following relationship for the magnetic field the expression for the theoretical H field is developed.

$$\tilde{H} = \frac{1}{\omega \mu} \left[\left(\frac{\partial E_z}{\partial y} - \frac{\partial E_y}{\partial z} \right) \hat{U}_x + \left(\frac{\partial E_z}{\partial x} \right) \hat{U}_y + \left(\frac{\partial E_y}{\partial x} \right) \hat{U}_z \right]. \tag{D-34}$$

Rearranging equation (D-13) and substituting the values for K' and R yields

$$E_{y} = \frac{E_{o}}{2\pi} \int \int \cos \frac{\pi_{x}'}{a} \frac{e^{-jkB} z(1-jkR)}{R^{3}} dx'dy'$$
(D-35)

$$E_{z} = \frac{E_{o}}{2\pi} \iint_{Aper} \cos \frac{\pi_{x}!}{a} \frac{e^{-jkB}(y! - y) (1 - jkR)}{R^{3}} dx'dy'$$
 (D-36)

where

$$R = \left[z^2 + (x - x')^2 + (y - y')^2\right]^{1/2}$$
 (D-37)

$$B = \left[\frac{x^{\frac{1}{2}}}{2 \ell_{H}} + \frac{y^{\frac{1}{2}}}{2 \ell_{E}} - R \right] , \qquad (D-38)$$

Taking the partial derivative of E_{z} with respect to y produces

$$\frac{\partial E_z}{\partial y} = \frac{E_o}{2\pi} \iint_{\text{Aper}} \cos \frac{\pi x'}{a} \frac{e^{-jkB}}{R^3} \left[(y-y')^2 \left(k^2 - \frac{3}{R^2} + j3\frac{k}{R} \right) + 1 - jkR \right] dx'dy'. \tag{D-39}$$

Similarly,

$$\frac{\partial E_{y}}{\partial z} = \frac{E_{o}}{2\pi} \int \int_{Aper} \cos \frac{\pi x'}{a} \frac{e^{-jkB}}{R^{3}} \left[z^{2} \left(k^{2} - \frac{3}{R^{2}} + \frac{j3k}{R} \right) + 1 - jkR \right] dx'dy' \qquad (D-40)$$

$$\frac{\partial E_{z}}{\partial x} = \frac{E_{o}}{2\pi} \iint_{Aper} \cos \frac{\pi x'}{a} \frac{e^{-jkB}}{R^{3}} (y-y') (x-x') \left(k^{2} - \frac{3}{R^{2}} + \frac{j3k}{R}\right) dx'dy' \quad (D-41)$$

$$\frac{\partial E_y}{\partial x} = \frac{E_o}{2\pi} \int \int \cos \frac{\pi x'}{a} \frac{e^{-jkB}}{F^3} z (x-x') \left(k^2 - \frac{3}{R^2} + \frac{j3k}{R}\right) dx'dy' . \quad (D-42)$$

From equation (D-24) the expression for the horizontal component of the magnetic field (H_{\bullet}) is

$$H_{x} = \frac{1}{\omega \mu} \left[\frac{\partial E_{z}}{\partial y} - \frac{\partial E_{y}}{\partial z} \right] . \tag{D-43}$$

Substituting equations (D-39) and (D-40) into equation (D-43) yields

$$H_{x} = \frac{jE_{o}}{2\pi\omega\mu} \left\{ \iint_{Aper} \cos \frac{\pi x'}{a} \frac{e^{-jkB}}{R^{3}} \left\{ \left(k^{2} - \frac{3}{R^{2}} + 3j\frac{k}{R} \right) (y-y')^{2} - z^{2} + \left(1-jkR \right) \right\} dx'dy' \right\} . \tag{D-44}$$

Where the vertical magnetic field (H_y) is

$$H_{v} = \frac{-j}{\omega \mu} \frac{\partial E_{z}}{\partial x} . \qquad (D-45)$$

Then,

$$H_{y} = \frac{jE_{o}}{2\pi\omega\mu} \int \int \cos \frac{\pi x'}{a} \frac{e^{-jkB}}{R^{3}} (y-y') (x-x') \left(k^{2} - \frac{3}{R^{2}} + \frac{j3k}{R}\right) dx'dy'. (D-46)$$

From equation (D-34) the radial magnetic field (${\rm H_{_{\rm Z}}}$) is

$$H_{z} = \frac{j}{\omega \mu} \left(\frac{\partial E_{y}}{\partial x} \right) \qquad (D-47)$$

Using equation (D-42) with (D-47) gives

$$H_{z} = \frac{jE_{o}}{2^{\pi\omega\mu}} \int \int \cos \frac{\pi x'}{a} \frac{e^{-jkB}}{R^{3}} z (x-x') \left(k^{2} - \frac{3}{R^{2}} + \frac{j3k}{R}\right) dx'dy' . \quad (D-48)$$

Equations (D-35), (D-36), (D-44), (D-46), and (D-48) represent the EM fields predicted by the aperture model.

APPENDIX E

EXPERIMENTAL DETERMINATION OF THE ELECTROMAGNETIC FIELD IN THE NEAR-ZONE OF A HIGH POWER UHF PYRAMIDAL HORN ANTENNA

This entire appendix appeared as a paper in the Proceedings of IEEE Southeastcon 1981, pages 503-507, Thomas H. Shumpert, Thomas A. Blalock and George R. Edlin.

EXPERIMENTAL DETERMINATION OF THE ELECTROMAGNETIC FIELD IN THE NEAR-ZONE OF A HIGH POWER UHF PYRAMIDAL HORN ANTENNA

ABSTRACT

The far zone characteristic (directive gain, main lobe beamwidth, side lobe structure, etc.) of pyramidal horn antennas are fairly well-known and, for the most part, easily predictable. However, as the observation point moves within the 'so-called" far zone boundary (often taken to be $2D^2/\lambda$), predicting the radiated field characteristics becomes much more difficult. This paper presents preliminary results of an intensive effort utilizing electrically small electric (dipoles) and magnetic (loops) probes to map the field in the region immediately in front of a high power UHF pyramidal horn antenna. The measured values are compared with values obtained from a theoretical model. Application of these data to electromagnetic hazard considerations are discussed.

INTRODUCTION

In the area of testing for electromagnetic radiation hazards (EMRH), electromagnetic compatibility (EMC), electromagnetic interference (EMI), accurate knowledge of the electromagnetic (EM) field into which the test item (or system) is to be immersed is of essential importance. Most common tests of these types include several apriori assumptions. First, the test item does not interfere (couple) with the mechanism for producing and maintaining the EM field environment, i.e., the radiating antenna structure, the parallel plate transmission line, the anechoic chamber walls, etc. Second, the impinging EM field is commonly assumed to be planar and uniform (true plane waves do not fall off with distance from the source). Third, the EM environment is accurately calibrated at various amplitude levels and predetermined frequencies of interest. Although the assumptions are usually dismissed as routine and well understood, the test engineer in EMRH, EMC, and EMI work is (or should be) all too familiar with the severe problems that arise upon violation of any of these simplifying statements. As expected, these assumptions apply quite well when the test item is sufficiently removed from the EM source (i.e., the test item is truly in the far zone of the EM source structure.) However, this "sufficient" separation is all too often compromised in actual test configurations. However, these compromises are not made out of ignorance or stupidity but are rather forced on the test engineer out of real necessity. Such considerations as limited real estate, limited radiating power, and large test items contribute significantly to this problem.

The purpose of this paper is to discuss measurement techniques and results of the application of these techniques to the situations in which one or more of these assumptions must be compromised. Specifically, this paper addresses the experimental determination of the EM field environment in the immediate region in front of a high power UHF pyramidal horn antenna.

UHF HORN CHARACTERISTIC

The specific horn antenna which is addressed in this paper is presently in use at the US Army Missile Command (MICOM) EM and Nuclear Effects facility

at Redstone Arsenal, AL. The horn is of a similar design to commercially available standard gain horns such as those supplied by the Narda Microwave Corporation or Scientific Atlanta. It is constructed of sheet aluminum with all seams formed with a heli-arc technique. The aperture dimensions are 2.44m by 1.8m and the overall feed-to-aperture dimension is approximately 4m. This antenna is designed to operate in the fundamental TE₀₁ mode radiation pattern over the UHF frequency range from 300 MHz to 550 MHz. The feed mechanism is coaxial-to-waveguide adapter with a "tear-shaped" stub for better broadband operation. The horn is mounted such that its center line is located approximately 4.6m (15 feet) above the ground, and it is oriented for vertical electric field polarization. The electric field pattern is the standard gain horn pattern with E and H plane 3dB beamwidths of approximately 22°. The nominal gain over the recommended band of operation is 18dB.

ANALYTICAL MODEL

For purposes of analyzing and comparing the measured IM fields with predicted values, a rather simple analytical model of this horn was developed for calculating the radiating properties. Figure 1 is a sketch of the horn geometry with the associated coordinate system.

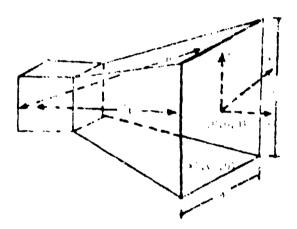


Figure 1. Geometry of UHF pyramidal horn antenna and associated dimensions and coordinate system.

The radiation model assumes a fairly simple aperture distribution as defined below

$$\tilde{E}(x',y') = U_y E_0 \cos(\frac{\pi x'}{a}) \exp[jk (\frac{x'^2}{2\ell_H} + \frac{y'^2}{2\ell_E})]$$
 (1)

This may be recognized as the ${\rm TE}_{01}$ mode amplitude distribution with a quadratic phase term. Several investigators^{1,2} have utilized this type of aperture function successfully. The radiated electric field intensity is then given by

$$\bar{E}(x,y,z) = \frac{1}{2^{\pi}} \nabla x \int \int [\hat{n}x\bar{E}(x',y')] \frac{e^{-jkR}}{R} dx'dy'$$
Aperture (2)

where

$$R = [(x-x')^2 + (y-y')^2 + z^2]^{1/2} . (3)$$

The evaluation of the curl and cross product operators yields the following

$$E_{\mathbf{x}}(\mathbf{x},\mathbf{y},\mathbf{z})=0 \tag{4}$$

$$E_{y}(x,y,z) = \int_{-a/2}^{a/2} \int_{-b/2}^{b/2} \cos(\frac{\pi x'}{a}) \exp(-jkb)z$$
.

$$(1-jkR) \frac{1}{R^3} dx'dy'$$
 (5)

$$E_z(x,y,z) = \int_{-a/2}^{a/2} \int_{-b/2}^{b/2} \cos(\frac{\pi x'}{a}) \exp(-jkb)$$
.

$$(y-y')$$
 $(1-jkR)$ $\frac{1}{R^3}$ $dx''dy'$ (6)

where

$$b = \frac{x^{+2}}{k} + \frac{y^{+2}}{k} - R$$
.

Substitution of equations (4)-(6) into Maxwell's equations yields the corresponding magnetic field components. Equations (4)-(6) and the corresponding ones for the magnetic field components were evaluated numerically. The results of these calculations and their comparisons to measured data are discussed in a later paragraph.

NEAR-FIELD MEASUREMENT TECHNIQUES

In order to map (measure) the EM field in the immediate vicinity of this UHF pyramidal horn, several pieces of experimental equipment were developed. Probably the most important development was the fabrication of electric and magnetic probes. These probes are electrically small, diode-loaded dipoles and loops. Their general design has been discussed in several National Bureau of Standards technical reports3,4. The specific probes utilized in this effort are discussed in great detail in another technical paper in this same meeting⁵. For this discussion, it is sufficient to describe them as electrically small, field-type selective, polarization selective, and of non-perturbing design and construction. The electric and magnetic probes were mounted on a moveable carriage which was computer controlled. Measurements from each of these probes was obtained at points in a symmetric volume located directly in front of the horn antenna. With reference to the coordinate system of the model, this volume occupied the space defined by the coordinates: $-1.5 \le x \le 1.5$, $-1.2 \le y \le 1.2$, $0 \le z \le 4.6$. Resolution of these spatial variations was 0.1m. All probe outputs were fed into the computer controlled data acquisition system, and the actual measured results, the corresponding physical locations, and the field component type and orientation were stored together on a magnetic disk. The moveable carriage was constructed almost entirely of polyvinyl chloride (PVC) pipes. The actual probes and associated circuit components were mounted on low-loss "G-10" dielectric sheets. All electrical leads from the probes to the monitoring voltmeters were high resistance (carbon-filled) lines to minimize their interaction with the field to be measured. This measurement technique produced a set of six fundamental quantities (the rms magnitude of Ex, Ey, E_z , H_x , H_y , H_z) at each of the 36,425 unique spatial locations contained within this previously defined test volume. Preliminary results gleaned from this large mass of data are presented in the following paragraph.

PRELIMINARY ANALYTICAL AND EXPERIMENTAL RESULTS

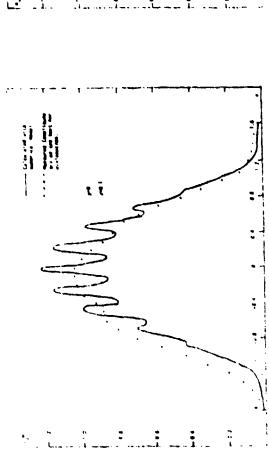
As was pointed out in the introductory paragraph, measurement or analytical prediction of the EM field in the region far removed from the source is quite straightforward. Near zone measurements and predictions usually are much more difficult. However, as a check on both the analytical and experimental techniques, far zone fields were predicted and measured. As expected, these predicted and measured fields were in very close agreement. Gain, E and H plane beamwidths, and $\frac{1}{r}$ spatial variations were very close to those previously predicted for horn antennas of this type. Consequently, all of the results discussed here will be near-zone analytical predictions and measurements. Furthermore all predictions and measurements were performed at 400 MHz.

Since it was assumed in the analytical model that the aperture distribution was of the TE_{01} amplitude variation, an interesting plane for measured observation is near the actual horn aperture (the z=0 plane). Figure 2 presents both the measured and predicted vertical electric field amplitude (rms $|E_y|$) as a function of x for a constant value of y and z (y=0, z=0.1). All analytical data has been amplitude normalized to match the average value of the measured data on the aperture center line (x=0, y=0). As can be seen,

this variation is indeed the expected cosinusoidal distribution. Figure 3 presents similar data for the horizontal magnetic field amplitude (rms |H_|). This also displays the expected cosinusoidal variation with x. Due to space limitations, the variations of these important aperture fields with y are not shown. However, both the analytical and measured curves for E, and H, vs. y show the expected uniformity of the assumed TE_{01} mode distribution. Figures 4 and 5 present identical data except that the plane of observation has been moved out to the z=4.6 plane. Here as expected, the plots of Ev and $H_{\mathbf{x}}$ show the typical far-zone variations, i.e., consinusoidal in both \mathbf{x} and y variations about the z-axis. Plots of the various other less significant field components as a function of x, y, or z have been made, but are not included due to space limitations. Figure 6 compares the measured vertical electric field on the hern axis (x=0, y=0) with both the predicted field and a superimposed $\frac{1}{2}$ fallout. All of these values are normalized to coincide at the point farthest from the horn (z=4.6). It is interesting and somewhat disturbing that the measured field does not exhibit the expected standing wave behavior as evidenced in the deep null in the analytical results near the aperture. This discrepancy continues to be an area of investigation. Finally, a plot of the z-directed wave impedance of the radiated EM field predicted by the numerical model is included. It demonstrates the expected asymptotic behavior toward 377 ohms as the observation point recedes from the aperture. However, its unexpected reversal of slope at points less than z ~ 0.75 suggest to these investigators that even the numerical model probably has significant deficiencies within a wavelength of the aperture.

CONCLUSIONS

The EM field in the near-zone of a high-power UHF pyramidal horn antenna has been determined both experimentally and analytically. The preliminary results suggest that the techniques employed and the resulting EM field measurements have yielded valuable data for characterizing the near field of the antenna. Much work still remains in analyzing the large quantity of experimental data. A technical report presenting the complete data analysis will be available in the near future.



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Figure 2. RMS magnitude of the vertical electric field component vs. x in the z=0.1 plane for y=0.

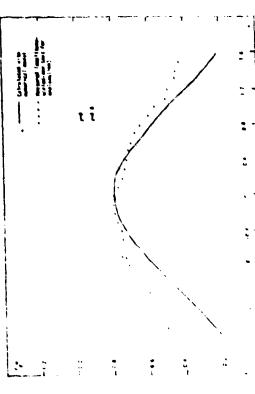


Figure 4. RMS magnitude of the vertical electric field component vs. x in the z=4.6 plane for y=0.

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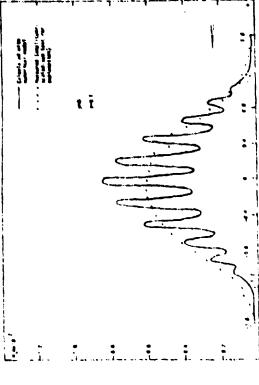


Figure 3. RMS magnitude of the horizontal magnetic field component vs. x in the z=0.1 plane for y=0.

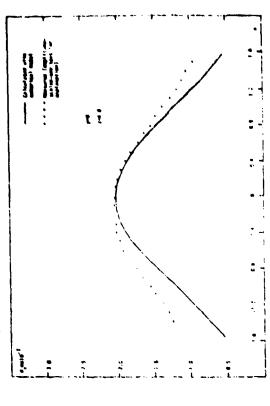


Figure 5. RMS magnitude of the horizontal magnetic field component vs. x in the z=4.6 plane for y=0.

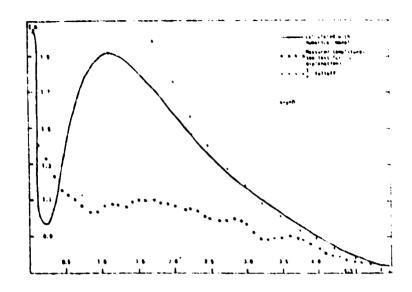


Figure 6. Analytical and experimental vertical electrical field component (rms magnitude) vs. z on the horn axis (x=y=0).

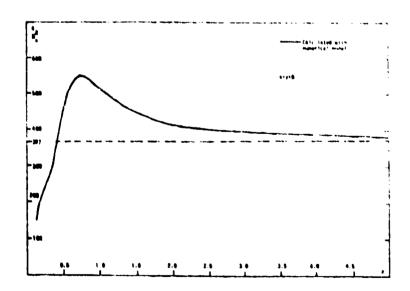


Figure 7. Z-directed wave impedance (analytical model only) vs. z.

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APPENDIX F

DERIVATION OF THE MULTIPOLE MODEL

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DERIVATION OF THE MULTIPOLE MODEL

It is desirable from the engineering standpoint to have simple models which can be used to predict E fields in the near-zone of standard gain horn antennas. Review of some selected references (Stratton, 162; Jackson, 136) suggests that complex field problems can be solved using infinite series expansion techniques. This method of field solution utilizes the superposition of the so-called dipole, quadrupole, octupole, and N pole terms, where n = 4 to ∞ . In addition, examination of the measured data demonstrates that the classical 1/r far field relationship does not hold close to the horn aperture. Thus the experimental data confirms the need for additional terms. Therefore, using the idea of the series expansion and truncating it at the third term, one has terms involving 1/r, 1/r² and 1/r³. By incorporating the appropriate constants with each of the terms, we have the dipole (A₁/r), quadrupole (A₂/r²) and octupole (A₃/r³) terms. Finally, incorporating the square root of the input power yields a model that can also relate the E field to the input power.

The assumed form for the multipole model is given as:

$$E = \left(\frac{A_1}{r} + \frac{A_2}{r^2} + \frac{A_3}{r^3}\right) \sqrt{r} . \qquad (F-1)$$

The constants A_1 , A_2 and A_3 are determined experimentally using measured data. The procedure is to choose three different distances and substitute these values along with the corresponding values of E into the generalized equation yielding three equations with three unknowns, which can then be easily solved simultaneously to give the proper values for A_1 , A_2 and A_3 . These values can then be normalized by dividing by the test power used in the original measurement. The final equation becomes:

$$E = \left(\frac{A_1}{r} + \frac{A_2}{r^2} + \frac{A_3}{r^3}\right) \left[P \text{ (Watts)}\right]^{1/2}$$
 (F-2)

where

E = E field (in V/m)

 A_1 , A_2 , A_3 = antenna constants

r = distance from the aperture to the field observation point
 (in meters)

P = power to the antenna (in Watts).

This simplified function can easily be programmed into a hand-held programmable calculator which will allow the determination of the near-zone E field at any distance from the aperture greater than $D^2/4\lambda$. The normalized equation for the 300-500 MHz test horn operating at 400 MHz is:

$$E = \left(\frac{44.95}{r} - \frac{46.17}{r^2} + \frac{1.46}{r^3}\right) P^{1/2} . {(F-3)}$$

This simplified equation can predict the field amplitude to approximately ±0.5 dB across the entire band of the horn antenna. If greater accuracy is required, the field must be carefully measured using calibrated probes. This model does not predict the E field at all frequencies. It is limited to the frequency range for which it is developed.

APPENDIX G

DEVELOPMENT OF THE ENGINEERING MODEL

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The engineering model is developed utilizing the following methodology.

The measured data is read at $2D^2/\lambda$, D^2/λ , $D^2/2\lambda$, and $D^2/4\lambda$. The equivalent far field gain is calculated and the effective gain plotted versus distance.

Observation of this graph suggests that the gain reduction in the near-sone would fit an exponential decay curve of the form:

$$G_{\eta\lambda} = G \left\{ 1 - \exp\left[-\alpha r/(2D^2/\lambda)\right] \right\}$$
 (G-1)

where

 $G_{n\lambda}$ = wavelength dependent near-zone corrected numerical gain

G = far field numerical gain

D = largest dimension of the aperture (in meters)

λ = wavelength (in meters)

a = gain attenuation constant.

For the horn utilized in this experiment, the approximate value for G is 58.8. Substituting into Eq. (G-1), one obtains

$$G_{\eta\lambda} = 58.8 \left\{ 1 - \exp \left[-\alpha r/(2D^2/\lambda) \right] \right\}$$
 (G-2)

when

$$r = D^2/\lambda$$
 , $G_{\eta\lambda} = 51.3$, $\alpha = 3.86$
 $r = D^2/2\lambda$, $G_{\eta\lambda} = 39.5$, $\alpha = 4.30$.

From the above calculations using the measured data and equation (G-2), one observes that α = 4 is a good choice for the gain attenuation constant. Putting this value into equation (G-2) produces:

$$G_{\eta\lambda} = 58.8 \left\{ 1 - \exp\left[-4r/(2D^2/\lambda)\right] \right\}$$
 (G-3)

This equation for the corrected near-field gain may be utilized in the standard far field formula. The following is the standard formula for the far field power density:

$$P_{d} = G_{T}P_{T}/4\pi r \tag{G-4}$$

where

 $P_d = power density (in Watts/m²)$

 $G_{\rm p}$ = numerical gain of the transmitting antenna

 $P_{_{\rm TP}}$ = power of the transmitter (in Watts)

r = distance from the antenna (in meters).

The following is the relationship between the E field and the power density:

$$E = (P_d^{377})^{1/2}$$
 (G-5)

where 377 is the free space wave impedance.

To develop the final form, the expression for $G_{\eta\lambda}$ is substituted into equation (G-4). This expression for P_d with the corrected gain is substituted into equation (G-5). This result gives the complete expression for E in the near-zone, as shown in equation (G-6):

$$E = \left\{ G_{m} \left[\frac{1 - \exp(-4r/(2D^{2}/\lambda))}{4\pi r^{2}} \right] P_{T}^{377} \right\}^{1/2} . \tag{G-6}$$

APPENDIX H

MEASURED FIELD DATA

(VERTICAL ELECTRIC FIELD ONLY)

APPENDIX H

MEASURED FIELD DATA (VERTICAL ELECTRIC FIELD ONLY)

This appendix contains a complete set of vertical E field data. The remaining five orthogonal components are not included due to the large volume of data. All components were measured with polarization selective, nonperturbing E and H field probes. These probes are described in detail in Appendix B. These electric and magnetic probes were mounted on a computer-controlled carriage which permitted a space of 3 meters x 2.4 meters x 4.8 meters to be covered with a resolution of 10 cm. The following coordinate system defines the rectangular parallelopiped mapped: $-1.5 \le X \le 1.5$, $-1.2 \le Y \le 1.2$, $0 \le Z \le 4.8$ (Figure 1).

The data are identified by the type E or H and by the specific component, i.e., radial, horizontal or vertical. At the beginning of each set of data, 'the position of the probe is given. For example, if the top probe position is 11 and one wishes to see the centerline data, the probe output to observe is Probe 4 which is at y = -1, as shown in Figure 2. Any desired vertical position relative to the center line of the horn can be identified by using Figure 2. Appendix C describes the probe positioner in detail.

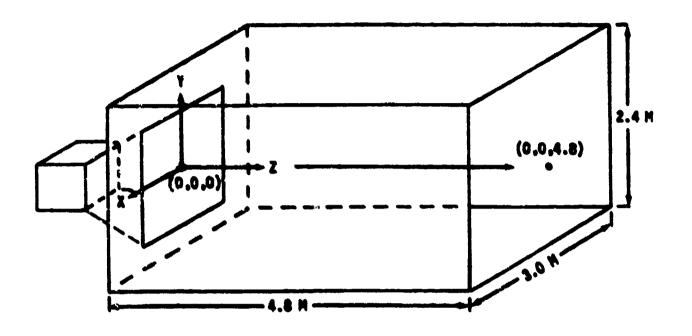


Figure 1. Volume of spatial E and H field measurements.

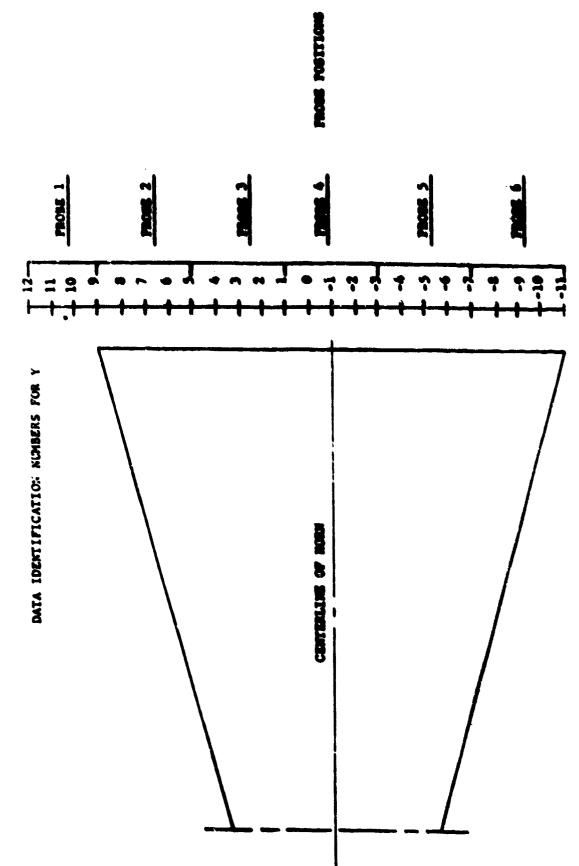


Figure 2. Vertical position of the magnetic and electric field probes relative to the horm of

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1.000 1.200	P1.362 8.559 10.362 10.373 12.	#1.31887 37.31887 33.5883 33.7754 33.5883 33.7754 33.7754 33.7754 33.7767 34.8777 36.3776 36.3	F1.116.5.286.5.84.9.916.6.7.9.3.3.5.5.4.4.9.5.28.3.3.5.5.4.9.9.3.3.5.5.3.3.3.5.5.3.3.3.5.5.3.3.3.5.5.3.3.3.5.5.3.3.3.5.5.3.3.3.3.5.5.3.3.3.5.5.3.3.3.5.5.3.3.3.5.5.3.3.3.5.5.3.3.3.5.5.3.3.3.5.5.3.3.3.5.5.3.3.3.5.5.3.3.3.5.5.3.3.3.5.5.3.3.3.5.5.3.3.3.5.5.3.3.3.5.5.3.3.3.5.5.3	P1 246 6 9 4 8 2 2 3 2 3 3 3 3 3 3 3 3 3 3 3 3 3 3 3	P. 1. 1. 1. 1. 1. 1. 1. 1. 1. 1. 1. 1. 1.	18.627775557 18.627775557 19.627775557 19.627775557 19.627775557 19.6277667 19.6277667 19.6277667 19.668 19.6

PRITCAL ELECTRIC FIELD STANTING COORDINATES PRINC 0 1.00 3.00 4.00 5.00 5.00	-3.00 -3.00 -3.00 -3.00 -3.00	12.00 8.00 4.00 0.00 -8.00	2 0.00 0.00 0.00 0.00 0.00			
7.000 0.100 0.200 0.300 0.400 0.500 0.500 0.500 0.500 1.000 1.200 1.300 1.400 1.500	-3.00 P(1) 5.981 8.868 11.142 12.307 13.456 16.734 16.575 18.876 21.587 21.588 21.587 21.589 21.587 21.589 19.783 18.545 18.545 18.154 17.328 18.154 17.328 18.154 17.328 18.154 17.328 18.154 17.328 18.154 17.328 18.154 17.328 18.154 17.328 18.154 17.328 18.154 17.328 18.154 17.328 18.154 17.328 18.154 17.328 18.154 17.328 18.155 18.	- E . 8 0 0 0 0 0 0 0 0 0 0 0 0 0 0 0 0 0 0	P(3) 55 47 67 75 75 75 75 75 75 75 75 75 75 75 75 75	P1433752443342.485444337752377754443373333333333333333333	P1.795551.942.3788.86.273.86.448.178.38.64.65.3.19.765551.98.648.75.36.448.75.36.448.178.38.648.33.98.62.23.19.626.3	# 1619
4.660	17.891	23.273	26.932	28.472	27.030	22.594

UFRICAL ELECTRIC FIELD STANTING CHURPINATES PRUE 0 1.80 2.00 3.00 4.00 5.00 6.00	-2.00 -2.00 -2.00 -2.00 -2.00	12.89 8.93 4.90 8.89 -4.08	7 0.00 0.00 0.00 0.00 0.00			
2 0.000 0.100 0.300 0.400 0.50	P10254 9.1946 11.126 14.038 12.1946 12.1946 12.1946 12.1946 12.1946 12.1946 12.1946 12.1947 12	F1275934884233333333333332228327263262222223344899993223333333333333333333333	P131 48.392 47.4633 50.8533 46.6533 46.6533 46.6533 46.6534 42.533 47.533 44.554 42.533 44.534 42.533 44.534 42.333 44.534 42.333 44.534 42.333 44.53	#1.343.57 #1.343.51 #1.343.51 #1.322.88 #1.323.89 #1.323.88	한 대한 19 19 19 19 19 19 19 19 19 19 19 19 19	P117.75113455450155000000000000000000000000000

WRITICAL ELECTRIC FIELD STARTIME COORDINATES PROBE \$ 1.00 2.00 2.00 4.00 5.00 6.00	-1.00 1.00 -1.00 -1.00 -1.00	1:1.00 9.00 4.00 -4.00 -4.00	0 . 00 9 . 0 9 . 0 9 . 0 9 . 0			
2 0.100 0.100 0.300 0.400 0.500 0.500 0.700 0.800 1.100 1.300 1.400 1.500 1.400 1.500 1.500 2.200 2.300 2.300 2.300 2.500	P(1) 5.157 9.194 11.561 13.120 14.160 17.721 17.721 17.721 17.722 21.565 21.765 21.766	P1.021 19.818 17.322 14.746 14.746 15.33.753 15.575 16.675 16.675 17.744	P1.464 19.464 19.464 19.464 19.464 19.464 19.464 19.464 19.534 19.664 19.534 19	PART 1 4 1 1 1 1 1 1 1 1 1 1 1 1 1 1 1 1 1	71.77.4.6.5.3.77.4.6.5.3.7.7.6.6.3.3.7.7.7.7.7.7.7.7.7.6.5.3.7.7.7.7.7.7.7.7.7.7.7.7.7.7.7.7.7.7	P.497.532 16447 17.532 16447 17.532 16447

VERTICAL ELECTRIC FIELD STARTING COGNICATES PROBE 0 2.00 3.00 4.00 5.00 6.00	X 9.00 9.00 9.00 9.00	Y 12.00 8.00 4.00 0.20 -4.10 -8.00	0.00 0.00 0.00 0.00			
2 0.000 0.108 0.200 0.300 0.400 0.500 0.500 0.500 0.700 0.700 1.000 1.300 1.400 1.500 1.500 1.500 2.500 2.500 2.500 2.500 2.500 2.500 3.100 3.100 3.100 3.100 3.100 4.200 4.300 4.400 4.50	P. 1.194 9. 1.194 13. 219 14. 245 14. 245 17. 365 18. 467 18. 467 18. 467 18. 467 18. 154 18. 154 18. 154 18. 154 18. 155 18. 17. 184 18. 239 17. 184 17. 18	P1.3374 41.3374 41.3374 527 53.374 53	## 131	15.056 14.5056 15.056 15.056 15.056 15.056 15.056 15.056 15.056 16.05	P. 1220 P. 122	**************************************

VERTICAL ELECTRIC FIELD STARTING COORDINATES PRODE P 1.00 2.00 3.00 4.00 5.00 5.00 5.00	X 1.00 1.00 1.00 1.00 1.00	12.00 9.00 4.00 0.00 -4.00 -8.00	7.00 9.00 0.00 0.00 0.00 0.00			
2	P.111 5.716 8.722 18.511 12.766 14.283 13.721 15.536 17.350 17.350 17.350 17.350 17.350 17.350 17.350 17.350 17.350 17.350 17.350 17.350 17.350 17.447 17.447 17.447 17.447 17.447 17.447 17.447 17.447 17.447 17.447 17.447	P.E.21 40.462 35.4649 33.198 33.553 33.653 33.653 33.653 33.779 31.666 33.2779 31.666 33.2779 31.666 5.672 26.694 26.694 27.849 26.694 26.551 24.282 25.283 27.282 24.282 24.282 25.283 27.282	77.36 1.17.36	P2.4945 41.4945 41.4945 41.4946 41.494	P.7.6813 P.7.6813 P.7.6813 P.7.6813 P.7.6813 P.7.6813 P.7.6813 P.7.6813 P.7.6813 P.7.6813 P.7.6813 P.7.694 P	P. 6815 P.

VERTICAL ELECTRIC FIELD STARTING COORDINATES PROPE 1 1.00 2.80 3.00 4.00 5.00 6.00	2.00 2.00 2.00 2.00 2.00 2.00	12.01 8.00 4.00 00 -4.00 -8.00	Z 0.00 0.00 0.00 0.00 0.00			
2 0.000 0.100 0.200 0.200 0.200 0.200 0.200 0.300 0.400 0.50	P11 5.628 8.119 9.13.341 13.727 14.728 13.341 13.727 14.728 14.729 15210 18.701 18.701 18.701 18.701 18.701 18.701 18.701 18.701 18.701 18.700 18.700 17.4	P. 121 36. 214 36. 214 36. 214 37. 198 38. 163 38. 163	P1.140 45240 45250 47824 47824 47824 44253 33314 48275 33314 48275 33314 33243 33317 34155 36354 37316 38317	Pt. 92-14 91. 92-14 91. 92-14 91. 92-14 92. 93-14 93. 93-14	P. 131 44.948 43.481 45.632 46.631 45.631 45.631 45.631 45.631 45.631 45.631 45.631 45.631 45.631 46.631 47.993 47.793 47	P. 681 165 165 165 165 165 165 165 165 165 16

VERTICAL ELECTRIC FIELD STARTING COURDINATES PROBE P 1.00 2.00 3.00 4.00 5.00 6.00	3.60 3.60 3.60 3.60 3.60	Y 0 0 9.00 4.00 0.00 -4.00 -8.00	0.70 0.00 0.00 0.00 0.70 0.70			
7.000 0.100 0.200 0.300 0.300 0.500 0.500 0.500 0.500 0.700 0.100 0.100 1.200	PI.533 9.1.289 11.899 11.899 11.899 12.44.975 11.899 12.44.9775 11.895 12.128 1	P1.4192 19.1198 19.119	PI 43.4618 44.4533 + 6.4533 +	P1.4931 44.6315 41.6315 41.6315 41.6316 41.631	917 0 6 4 9 1 1 4 1 5 1 7 9 8 8 7 1 1 4 1 5 1 7 9 8 8 7 1 1 4 1	P. 1913 161

VERTICAL-ELECTRIC FIELD STANTING COORDINATES PRUBE 1 .00 2.00 4.00 5.00 6.00	X 4.00 4.00 4.00 4.00 4.00	12.00 8.00 4.00 -4.00 -4.0	Z Q.00 Q.00 Q.00 Q.00 Q.00			
Z 0.000 0.100 0.200 0.200 0.300 0.400 0.50	\$\\\^{11} \\ \\^{11} \\\^{11} \\ \\^{11} \\ \\^{11} \\ \\^{11} \\ \\^{11} \\ \\^{11} \\	P121 35.744 21.841 38.1	P(3) 40.565 41.657 41.657 41.657 42.345 43.475 43.4	P. (40.6738) P. (40.6739) P. (4	P\$ 332 332 333 333 333 334 7 133 9 6 3 6 6 2 2 2 2 2 2 2 2 2 2 2 2 2 2 2 2	P. 181 7 1 2 2 3 3 4 4 6 9 8 2 5 4 5 2 5 2 5 2 5 2 5 2 5 2 5 2 5 2 5

VERTICAL ELECTRIC FIELD STARTING COOMDINATES PROME € 1.00 2.00 3.00 4.00 5.00 6.00	X 00 5.00 5.00 5.00 5.00 5.00	12.69 3.89 4.00 8.00 -4.00 -8.00	7.00 0.00 0.00 0.00 0.00			
7.000000000000000000000000000000000000	P111 4.294 7.248 8.464 9.493 10.364 11.561 12.393 13.428 13.478 14.283 15.415 16.853 17.17.525 16.853 17.012 17.012 17.013 16.734 16.734 16.734 16.734 16.736 15.325 15.495 15.495 15.495 15.495 15.495 15.495 15.577 16.857 16.857 16.857 16.857 16.857 16.857	P121 30.591 28.545 28.545 28.545 28.545 29.344 29.345 29.345 29.345 29.345 29.345 29.345 29.345 29.345 29.345 20.888 21.221.221.221.221.221.221.221.221.221.	PI 31 37 16 31 37 16 37	P1 308 3 99 3 6 8 3 1 5 4 7 1 2 3 2 3 3 3 3 3 3 3 3 3 3 3 3 3 3 3 3	77.11289.964.232.465.0652.07.11289.765.0652.07.1233.332.1.4888.332.1.4888.3322.1.569.09.19.19.19.19.19.19.19.19.19.19.19.19.19	P.318.235.64.13.88.22.78.63.88.78.63.88.78.63.88.78.78.63.88.88.78.78.78.78.78.78.78.78.78.78.78.

VERTICAL ELECTRIC FIELD STARTING COURDINATES PRURE 8 1.90 2.00 3.00 4.00 5.00 6.00	X 6.00 6.00 6.00 6.00	12.00 Fi.00 Fi.00 -4.00 -4.10 -8.00	2 0.80 8.80 9.00 0.00 0.00			
Z	P(11 4.115 6.596 7.989 8.636 7.785 11.184 11.811 12.357 11.174 15.857 16.376 16.377 16.377 16.377 16.389 17.299 17.299 17	P121 30.175 27.439 26.498 26.498 26.413 27.919 26.413 27.919 26.413 27.919 26.413 27.919 27.253 26.483 25.180 21.740 21.488 21.488 21.488 21.488 21.488 21.488 21.488 21.488 21.488 21.488 21.488 21.488 21.111 21.111 21.138 21.111 21.138 21.111 21.138 21.138 21.111 21.138 21.138	P131 74.289 74.289 74.715 74.715 75.4318 75.44.518 75.44.518 75.25 76.25	FU 5243 7 149 331 7 149 33	PISE 1 1 3 2 2 2 2 2 2 2 2 2 2 2 2 2 2 2 2 2	P(6430827, 6505) 27, 6505 25,

VERTICAL ELECTRIC Starting Coordina Pri	FIELD TES 1.00 7.00 2.00 7.00 3.00 7.00 4.00 7.00 6.00 7.00	Y 12,04 8,90 4,00 0,00 -4,80 -8,80	0.00 0.00 0.00 0.00 0.00			
Z 0.1:23 0.23 0.34 0.50 0.50 0.50 0.50 0.50 0.50 0.50 0.5	10	P(2) 24.819 23.273 22.918 23.498 22.546 22.583 23.759 24.977 24.977 24.977 24.977 24.977 24.977 24.977 24.977 24.977 24.977 24.977 24.977 24.977 24.977 28.179 19.879 19.879 19.879 19.879 19.464 19.575 19.728 19.728 19.728 19.728 19.728 19.728 19.728 19.728	P(3) 28.252 30.047 31.593 30.600 31.565 27.538 27.5	P. 141 277. 126 281. 197 261. 1839 261. 622 261. 623 261. 624 261.	P.1798 1.7798 1.7798 1.7798 1.7799 1.	P. (6) 221.7 222.6321 221.5552 222.632 221.5552 221.632 221.788 221.788 221.554 221.55

PERITCAL ELECTRIC FIELD STARTING COURDINATES PROBE \$ 1.00 2.00 3.60 4.00 5.00	X 8.84 8.00 8.11 8.11	12.00 8.00 4.00 4.00	Z 0.00 0.00 0.60 0.00			
8.00	8,00	8.90	1,11			
7 0.000 0.100 0.300 0.400 0.500 0.500 0.700 0.700 1.400 1.500 1.500 1.500 1.500 2.100 2.100 2.700 2.500 2.700 2.500 2.700 3.400 3.500 3.400 3.500 3.400 4.100 4.100 4.100 4.200	PU11 2.717 6.421 7.252 6.970 7.945 9.748 9.876 11.853 13.383 13.383 13.424 13.425 14.425 14.425 12.228 12.393 12.559 12.559 12.559 12.559 12.559 12.559 12.559	P[2] 20.142 18.975 19.502 20.366 17.916 21.047 17.540 21.763 21.763 21.000 21.7640 21.765 19.665 19.665 19.615 18.323 18.428 19.665 17.477 17.128 17.428 17.428 18.323 18.361 17.438 18.361 18.361 18.361 18.361 18.361 18.361 18.361 18.361 18.361	PI 31 22.845 23.469 25.858 27.489 26.479 26.489 27.489 26.853 26.967 24.489 24.485 23.868 23.488 23.	P1.41 23.122 22.577 21.868 22.577 21.433 24.410 25.477 26.477 27.254 27.254 28.609 28.335 28.369 28.335 28.369 28.335 28.348 26.778 26.778 26.975 26.975 26.374 26.378 26.379 26.379 26.379 26.379 26.379 26.379 26.379 26.379 27.379 28.369 27.379 28.	P (5) 5 20 20 20 20 20 20 20 20 20 20 20 20 20	P(6) 23-8-19-19-19-19-19-19-19-19-19-19-19-19-19-
4.588 4.680	13.875 14.128	18.875	22.387	23.085	21.920	18.254

VERTICAL ELECTRIC FIELD STAKTING COURDINATES PRINE \$ 1.00 2.00 3.00 4.00 5.00 6.00	X 9.80 9.80 9.80 9.80	12.00 8.00 4.00 0.00 -4.00 -8.00	2 0.00 0.00 0.00 0.00			
2 0.000 0.20	P111 1.877 3.530 5.420 5.544 7.514 9.236 9.236 9.236 11.875 12.228 12.557 11.978 11.436 11.436 11.436 11.561 11.56	P123 13.893 13.732 14.853 15.324 15.324 16.354 18.361 18.361 18.361 18.361 18.361 18.361 18.361 18.361 18.361 18.361 18.361 18.361 18.361 18.361 18.361 18.361 16.1781 16.1781 16.739 16.739 17.862 18.739 17.862 17.862 17.862 17.862 17.878 17.878 17.878 17.878 17.878 17.878 17.878 17.878 17.878 17.878 17.878 17.878 17.781 17.781 17.781 17.781 17.781 17.781 17.781	P131 14.779 14.778 17.726 21.457 20.770 21.313 22.280 21.807 21.807 21.807 21.807 21.807 21.807 21.807 21.807 21.807 21.807 21.807 21.809 21.8	#141	P. [3] 13.41 27.22 13.41 27.23 20.737 20.737 21.738 21.738 21.738 21.738 21.738 21.343	P. (63) 13. 7349 14. 8019 15. 8289 14. 8019 15. 8289 16. 8289 16. 8289 16. 8289 16. 8289 16. 8289 16. 8289 16. 8289 16. 8289 16. 8289 16. 8289 16. 8289 16. 8289 16. 8289 16. 8289 16. 8289 16. 8289 16. 8289 16. 8289 16. 8289 17. 8299 17.

VERTICAL ELECTRIC FIFLD STANTING COORDINATES PROPE 1.00 2.00 3.00 4.00 5.00 6.00	-15.00 -15.00 -15.00 -15.00 -15.00	9.00 5.00 1.90 -3.00 -7.00 -11.00	0.00 0.00 0.00 0.00 0.00			
2 0.000 0.200 0.300 0.400 0.500 0.500 0.500 0.500 0.900 1.100 1.50	P111 2.126 3.575 5.185 5.187 5.628 5.589 7.469 8.765 9.876 10.342 11.100 12.661 12.681 12.681 12.681 12.897 13.855 12.898 11.858 11.811 11.858 11.811 11.858 11.811 11.859 11.978 11.978 12.269 11.978 12.269 11.978 12.269 11.978 12.313 11.978 12.328 11.352	P121 14.891 14.173 15.168 16.234 16.316 18.933 17.743 18.893 17.824 18.893 17.824 18.893 17.824 16.388 16.388 16.388 16.388 16.388 16.388 16.388 16.388 16.388 16.388 16.388 16.388 17.431 16.388 17.798 16.483 17.798 17.798 17.798 17.798 17.798 17.798 17.798	P(31 15.724 17.756 12.724 17.756 12.794 21.672 21.672 21.672 21.672 21.798	P141 16.433 17.4558 17.558 17.558 17.865 17.865 17.865 17.865 17.865 17.865 17.865 17.865 17.865 17.865 17.865 18.558 17.565 18.558 18.	P1.240 15.847 16.5157 20.328 21.781 22.4950 21.781 22.4950 21.781 21.781 21.781 21.781 21.781 21.781 21.783	P (6) 476 14. 138 15. 14. 158 15. 157 157 157 157 157 157 157 157 157 157

VERTICAL ELECTRIC FIELD STARTING COURDINATES PRIVE 9 1.00 2.00 3.00 4.00 5.00 6.00	7.00 7.00 7.00 7.00 7.00 7.00	12.88 9.80 4.80 1.80 -4.80 -8.81	0.00 0.00 0.00 0.00 0.00			
7 0.000 0.100 0.300 0.410 0.500 0.500 0.500 0.500 0.500 1.000 1.300 1.400 1.500 1.500 1.500 2.300 2.300 2.300 2.500 2.500 2.500 2.500 3.400 3.500 3.500 3.500 3.500 3.500 3.500 4.100 4.200 4.100 4.300 4.100 4.500 4.500 4.500 4.500 4.500	P(1) 1.787 5.875 5.828 5.755 5.428 8.765 9.368 11.142 12.645 12.645 12.645 12.645 12.645 11.752 12.848 11.875 11.8	P(2) 14.970 14.973 15.208 15.975 18.975 18.975 18.975 18.975 18.975 18.975 18.975 18.975 18.975 18.975 18.975 18.975 18.975 18.975 16.507 16.507 16.507 16.308 16.308 17.807 17.8	P: 31 15.724 17.407 20.160 22.630 22.630 22.630 22.630 22.733 21.457 21.457 22.387 22.	71.797 16.797 16.490 17.2443 16.490 17.2443 16.490 17.2413 16.397 21.4785 22.4798 22.4798 23.489 24.798 25.798 26.184 26.537 27.763 27.	F1.356 14.357 15.577 16.727 11.708 11	P(6) 14.243 14.243 15.338 15.368 15.368 16.5715 16.5715 17.3238 16.3210 16.3210 16.499 16.499 16.499 16.499 16.499 16.499 17.499 18.475 17.498 18.498 17.498 17.498 17.498 18.498 17.498 17.498 18.498 17.498 18.498 17.498 18.498 17.498 18.498 17.498 18.498 17.498 17.498 18.498 17.498 18.498 17.498 18.498 17.498 18.498 17.498 18.498 17.498 18.498 17.498 18.498 17.498 18.498 17.498 18.498 17.498 18.498 17.498 17.498 18.498 17.498 18.498 17.498

PERTICAL ELECTRIC FIELD STARTING COORDINATES PRIME \$ 1.00 2.00 3.00 4.00 5.00 6.00	X 10.00 10.00 10.00 10.00	12.00 8.00 9.00 9.00 -4.00 -8.00	2 0.00 0.00 0.00 0.00 0.00			
2 0.000 8.100 0.20	P(1) 1.649 2.717 4.740 4.740 4.740 5.274 6.245 7.663 10.575 11.720 11.663 11.720 10.674 11.720	P121 11.051 12.075 13.007 13.007 13.007 14.173 15.366 16.310 15.232 16.732 16.339 16.339 16.339 16.153 15.007 17.007 17.007 17.007 17.007 17.007 17.007 16.349 16.556 17.007 17.007 17.007 16.349 16.508	P131 19.667 13.625 16.621 18.625 19.621 19.623 19.653 19.653 19.653 19.653 19.653 19.653 19.653 19.653 19.74 19.943 20.269 20.846 21.27 21.241 21.457 21.241 21.457 21.241	#11.673 11.673 11.673 12.874 13.574 13.574 16.730 17.377 16.730 17.377 16.730 17.377 16.730 17.377 1	P. 151 11.250 14.277 16.260 17.157 18.261 17.157 18.262 18.262 19.569 19.563 19.563 19.563 19.570 20.763 20	P(61) 10.444 10.752 11.756 12.276 12.664 13.3787 14.655 15.429 16.425 16.425 16.423 16.623 16.623 16.623

MERTICAL ELECTRIC FIELD STRATING COORDINATES PRINT ? 1.00 2.00 4.00 5.00 6.00	11.00 11.00 11.00 11.00 11.00 11.00	12.00 9.09 1.00 0.00 4.00 8.09	0.00 0.00 0.00 0.00 0.00			
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VERTICAL ELECTRIC FIELD STARTING COURSINATES PROSE \$ 1.00 2.00 3.00 4.00 6.00	12.00 12.00 12.00 12.00 12.00	Y 12.00 8.99 4.00 0.00 -4.00 -8.00	2 1:10 1:10 1:11 1:11			
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VERTICAL FLECTRIC FIELD STARTING COURDINATES PRUBE # 1.00 2.00 3.00 4.10 5.00 6.00	X 14.90 14.90 14.90 14.90 14.96	12.00 8.00 4.05 0.00 -4.00	Z 9.00 0.00 0.00 0.00 0.00			
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ELECTRIC FIELD COORDINATES PROBE (1.69 2.90 3.00 4.00 5.00 6.00	15.00 15.00 15.00 15.00 15.00	12.00 8.09 4.00 -4.00 -8.00	0.00 0.00 0.00 0.00 0.00			
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VERTICAL STAKTING

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